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SU CAVO

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# Contents

<b>1</b>	<b>Introduction</b>	<b>5</b>
1.1	Abbreviations . . . . .	5
1.2	Definitions . . . . .	6
<b>2</b>	<b>Systems Architecture</b>	<b>9</b>
<b>3</b>	<b>Input Processing</b>	<b>13</b>
3.1	Mode Adaptation . . . . .	13
3.2	Stream adaptation . . . . .	15
<b>4</b>	<b>Bit Interleaved Coding and Modulation</b>	<b>17</b>
4.1	FEC Encoding . . . . .	18
4.2	Mapping bits onto constellations . . . . .	20
<b>5</b>	<b>Data Slices and L1 signalling</b>	<b>27</b>
5.1	Data Slice Packet Generation . . . . .	27
5.2	C2-System: Layer 1 part 2 signalling . . . . .	29
5.3	T2 system: L1 signalling . . . . .	35
<b>6</b>	<b>Frame Builder and OFDM generation</b>	<b>37</b>
6.1	C2 vs T2 Frame structure . . . . .	37
6.2	OFDM generation . . . . .	39
<b>7</b>	<b>Channel Model</b>	<b>47</b>
7.1	HFC Channel Model . . . . .	47
7.2	Wireless Channel . . . . .	53
<b>8</b>	<b>Conclusions</b>	<b>61</b>
	<b>Bibliography</b>	<b>63</b>



# Chapter 1

## Introduction

The second generation of DVB physical layer standard family (DVB-x2) consists of three transmission systems for three different transmission media: DVB-S2 (satellite), DVB-T2 (terrestrial) and DVB-C2 (cable).

The first part of this thesis makes a comparison between the two systems DVB-T2 and DVB-C2, describing the architectures of the two systems and highlighting the main differences and similarities. It specifies the channel coding and modulation, the lower signaling protocol system, the Frame Builder and the OFDM generation.

The second part defines the channel model and gives a description of the HFC and the wireless channel model.

### 1.1 Abbreviations

**ACE** = Active Constellation extension

**BCH** = Bose and Ray-Chaudhuri code

**BER** = Bit error ratio

**BICM** = Bit Interleaved Coding and Modulation

**DFL** = Data Field Length

**DNP** = Delete Null Packet

**CRC** = Cyclic Redundancy Check

**FEC** = Forward Error Correction

**FEF** = Future Extension Frame

**GSE** = Generic Stream Encapsulation

**HEM** = High Efficiency Mode

**HFC** = Hybrid Fibre Coaxial

**IM** = Intermodulation

**ISI** = Intersymbol Interference

**LDPC** = Low-Density Parity-Check code

**LTI** = Linear Time-Invariant

**MISO** = Multiple Input Single Output

**NM** = Normal Mode

**OFDM** = Orthogonal Frequency-Division Multiplexing

**PAL** = Phase Alternating Line

**PAPR** = Peak to average power ratio

**PLP** = Physical Layer Pipe

**PSD** = Power Spectral Density

**QAM** = Quadrature Amplitude Modulation

**RMS** = Root Mean Square

**SNR** = Signal-to-noise ratio

**TI** = Time Interleaver

**TS** = Transport Stream

**UP** = User Packet

**XFECFrame** = FECFrame mapped onto QAM constellations

## 1.2 Definitions

**Bit error rate** = the number of bit errors divided by the total number of transferred bits.

**Channel raster** = difference of frequencies, e.g. the frequency difference between two adjacent channels.

**Code rate** = the code rate of a forward error correction code is the ratio between the number of uncoded bits and the total number of coded bits (data and redundancy).

**FEF part** = part of the super-frame between two T2-frames which contains FEFs.

**Generic Stream Encapsulation protocol** = a data link layer protocol defined by DVB ([2]).

**Guard Interval** = interval used to ensure that distinct transmissions do not interfere with one another, e.g. distinct transmissions of the same user as in OFDM. The purpose of the guard interval is to introduce immunity to propagation delays, echoes and reflections, to which digital transmissions are usually very sensitive.

**In-band signalling** = metadata and control information in the same band or channel used for data.

**Parity Bit** = a bit that is added to ensure that the number of bits with the value one in a set of bits is even or odd. Even parity is the special case of CRC and of the Bit Interleaver.

**Transport Stream** = a standard format for transmission and storage of audio, video, and data. Transport stream specifies a container format encapsulating packetized elementary streams, with error correction and stream synchronization features for maintaining transmission integrity when the signal is degraded. ([1]).





## Chapter 2

# Systems Architecture

The block diagrams' DVB-T2 and DVB-C2 systems are given from figure 2.1 to figure 2.5.

The system (C2 or T2) input(s) may be one or more MPEG-2 Transport Stream(s) (TS, see definitions) and/or one or more Generic Stream(s) data based on the Generic Stream Encapsulation protocol (GSE, see definitions).

The Input Pre-Processor, which is not part of the C2/T2 system, provides the creation of the individual Physical Layer Pipes (PLPs), that are logical channels containing the TS or the data based on GSE. It may include a Service splitter or de-multiplexer for Transport Streams (TS) for separating the services into the T2/C2 system inputs, which are one or more logical data streams.

The PLP consists of a PLP Id, a 8-bit field which identifies uniquely a PLP within a C2/T2 transmission signal: after decoding this header the receiver is able to decide whether it has to decode the following data packet. Data packets that do not belong to the requested PLP are ignored so that the effective receiver bit-rate as well as the related processing power decrease significantly. Therefore, error-correction coding and interleaving are applied separately to each PLP. This allows for the implementation of service-specific live. Specifically for the C2-system this is important for giving a high service quality to all reception points in the cable network. In fact the cable characteristics like distance, number of amplifiers, quality of in-house installations, affect the signal quality that can vary significantly between all the connections.

The system output is typically a single signal to be transmitted on a single RF channel.

Optionally, the T2-system can generate a second set of output signals, to be conveyed to a second set of antennas in what is called Multiple Input Single Output (MISO) transmission mode: a description of that is given in the section of OFDM generation (chapter 6).

An important constraint over the duration of one physical-layer frame

(T2/C2-frame) is that the total input data capacity (in terms of cell throughput, following null-packet deletion, if applicable, and after coding and modulation) shall not exceed the T2/C2 available capacity (in terms of data cells, constant in time) of the T2/C2-frame for the current frame parameters. The systems adopt different solutions to achieve that constraint.

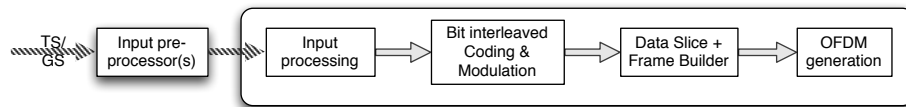
In the T2-system the PLPs are arranged in groups, each originated from a single, constant bit-rate, statistically-multiplexed source. Therefore each group will always use the same modulation, coding and interleaving depth.

In the C2-system one or more PLPs are arranged in a group of PLPs and one or more of such groups of PLPs form a Data Slice. Every C2-System contains one or more Data Slices.

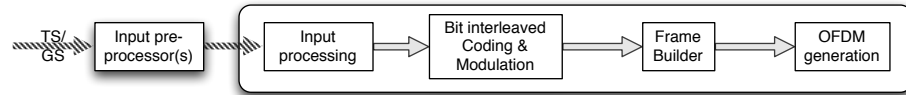
Each group of PLPs may contain one common PLP, but it isn't necessary; when the DVB-T2/C2 signal carries a single PLP there is no common PLP.

The total System Architecture is divided into the following processes:

1. Input Processing;
2. Bit Interleaved Coding and Modulation (BICM);
3. Frame generation (by the Frame Builder)
4. OFDM generation.

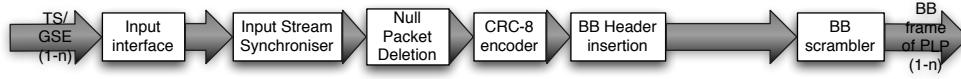


(a) System Overview DVB-C2

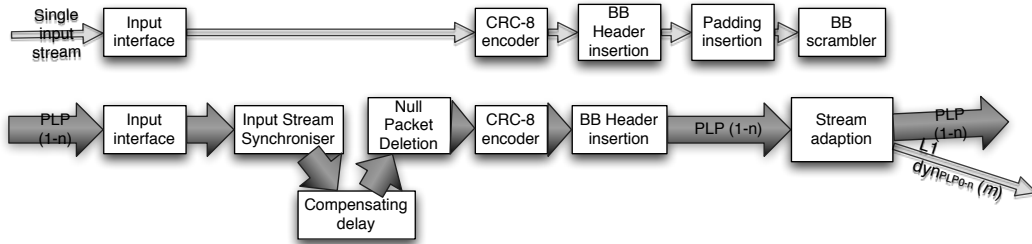


(b) System Overview DVB-T2

Figure 2.1: Comparison of the systems' architecture.

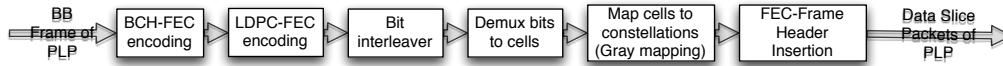


(a) Input Processing DVB-C2

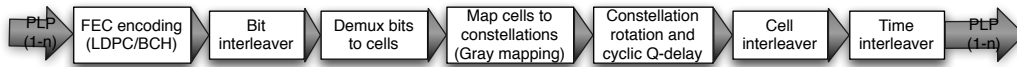


(b) Input Processing DVB-T2: single PLP (first chain) and multiple PLP (second chain)

Figure 2.2: The Input Processing of the two systems: note that Compensating Delay is an additional block of the T2-system.

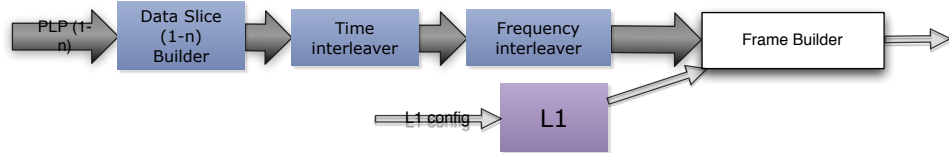


(a) BICM DVB-C2

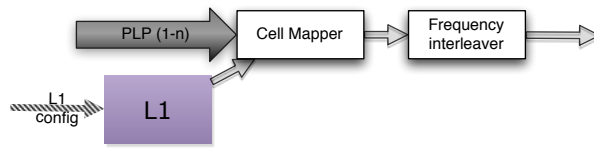


(b) BICM DVB-T2

Figure 2.3: The Bit Interleaved Coding and Modulation of the two systems: note that constellation rotation, cell and time interleaver are additional blocks of the T2-system, FEC-frame header insertion is an additional block of the C2-system.



(a) Frame Builder and Data Slice Builder DVB-C2

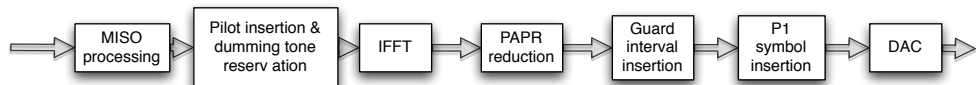


(b) Frame Builder DVB-T2

Figure 2.4: Frame generation of the two systems.



(a) OFDM generation DVB-C2



(b) OFDM generation DVB-T2

Figure 2.5: OFDM generation of the two systems: DVB-T2 has many additional blocks, as the MISO processing, the PAPR reduction and P1 symbol insertion.

## Chapter 3

# Input Processing

### 3.1 Mode Adaptation

As mentioned in the System Architecture the input of the T2/C2 system consists of logical data streams, each of them carried by an assigned PLP. The first part of Input Processing is represented by the Mode Adaption which, operating separately on the contents of each PLP, slices the input data into data fields, which at the end will form baseband frames (BBFrame). The mode adaptation module comprises the input interface, followed by three optional sub-systems (the input stream synchronizer, the Null Packet deletion unit and the CRC-8 encoder) and then finishes by slicing the incoming data stream into data fields and inserting the baseband header (BBHeader) at the start of each data field.

#### Input Interface

The input interface sub-system maps the input into internal logical bit-format. This operation is applied separately for each single PLP.

Generally the input is mapped into data fields composed of DFL bits (Data Field Length): this number can vary from 0 to a value depending on the chosen LPDC code. The 10-byte BBHeader is appended to the front of the data field. The user packets (UP) are allocated into data fields according to the capacity of the last ones; in this way different UPs can be broken in subsequent data fields or allocated in a single data field.

The principal difference in the Input Interface of the C2 and T2-system consists in the data field capacity (DFL). While in the first it's always equal to  $K_{bch}-80$ , in the last one this is true only when the in-band signalling is not used, otherwise it's smaller ( $K_{bch}$  is the number of bits of BCH uncoded Block, see the BICM section). When the value of  $DFL < K_{bch}-80$  both systems support a padding field to complete the LPDC/BCH code block capacity; in the T2-system a padding field is also applicable for transmit in-band

signalling.

### **Input Stream Synchronizer**

The Input Stream Synchronizer provides a mechanism to regenerate, at the receiver, the clock of the Transport Stream (or packetized Generic Stream), in order to guarantee end-to-end constant bit rates and delays of the user information. This is an optional block, except that it is always used for PLPs carrying transport streams where the number of FEC blocks per C2/T2 frame may vary.

In the C2-system the synchronizer works on a single PLP travelling in different Data Slices; in the T2-system it allows synchronization of multiple input streams travelling in independent PLPs. So that the reference clock and the counter of the input stream synchronizers are the same.

### **Compensating Delay**

In the T2 system there is an additional block, the Compensating Delay, required at the receiver in order to coordinate the delays of different streams. The data PLPs and the corresponding common PLP may have different parameters of the time interleaving and the frame interval. At the transmitter the information of the data streams is split in the data PLPs and in the common PLP; for correctly receive the data streams, the receiver has to read the data PLPs and the common PLP at the same time. In this way the partial transport streams at the output of the dejitter buffers (a sort of buffer responsible to restore synchronization) for data and common PLPs would be essentially co-timed.

### **Null Packet Deletion**

This option allows the system to delete the null packets made at the Synchronizer to guarantee end-to-end constant bit rates and delays. The packets deleted at the transmitter will be restored at the receiver in the exact place where they were originally (so that synchronization is given) by the insertion at the transmitter of a DNP field (Deleted Null-Packets, 1 byte) in place of the null packet. The DNP represents the number of the null packets deleted and will be reset when it reaches the maximum allowed value 255, with the transmission of the next packet in any case, also if the following packet is again a null packet.

### **CRC-8 encoding**

CRC-8 encoding is applied for error detection at the User Packet (UP) level; it's applicable only in the normal mode (see the next paragraph). This makes CRC words by the implementation of Cyclic Redundancy Check code

that allows the detection of transmission errors at the receiver side. The CRC-device calculates a short, fixed-length binary sequence of 8 symbols (the CRC word), for each block of data to be sent or stored and appends it to the data, forming a codeword. At the receiver a device compares the codeword received with a set of codewords expected (the receiver knows the rule for the calculation of the codeword and so the set of possible codewords). If the codeword doesn't belong to this set, then the block contains a data error and the device may take corrective action such as reading or requesting the block be sent again, otherwise the data is assumed to be error-free.

### **Baseband Header insertion**

The BBHeader represents 10 bytes that are inserted in front of the baseband data field in order to describe the format of the data field. It's the same for the two systems and it comprises two forms: the normal mode (NM) and the high efficiency mode (HEM). These modes in combinations with the Input stream format (Generic Packetized Stream, Transport Stream, Generic Continuous Stream and Generic Encapsulated Stream) define the mode adaptation processing.

The first two bytes represent the MATYPE, that describes the input stream format and the type of the Mode Adaptation. The other bytes hold other information like as the user packet length, the data field length etc..

## **3.2 Stream adaptation**

The output of the mode adaptation is a BBHEADER followed by a DATA FIELD, the stream adaptation provides three processes: scheduling, padding and scrambling.

### **Scheduler**

The C2-scheduler works together the Data Slice builder to decide which Data Slices of the final C2 System will carry data belonging to which PLPs. At the end of this process it generates L1-part2 signalling information (see chapter 5) and it defines the composition of the Data Slice and C2 Frame structure.

The T2-scheduler has a similar function in order to generate the required L1 dynamic signalling information (see chapter 5). Like the C2-scheduler it defines the composition of the frame structure.

### **Padding**

As introduced in the Interface's section, padding may be applied when the BBFRAME isn't full, or when an integer number of UPs has to be allocated

in a BBFRAME. Intuitively, the number of padding bits is  $K_{bch}$ -DFL-80, so that the LPDC/BCH code block capacity is completely filled ( $K_{bch}$  bits at the BBFRAME). In the T2-system the PADDING field may also be used to carry in-band signalling.

### **BB scrambling**

The last block of the input processing is the BB scrambler, equally for the two systems it randomizes the sequence of the BBFRAME to reduce the energy dispersal.



## Chapter 4

# Bit Interleaved Coding and Modulation

One of the main characteristics of the second digital video broadcasting is the introduction of the LDPC and BCH encoding, that gives a great improvement in the forward error correction encoding (FEC).

The following example gives an idea of this statement (see also [4]). The 9/10 code rate of DVB-C2 is able to correct bit streams with bit-error rates of several percent measured at the input of the FEC decoder. In contrast, the Reed Solomon code applied for DVB-C, which has a similar effective code rate, only tolerates a maximum bit-error rate of  $2 \cdot 10^{-4}$  to reach the goal of quasi error-free reception corresponding to one erroneous event per hour.

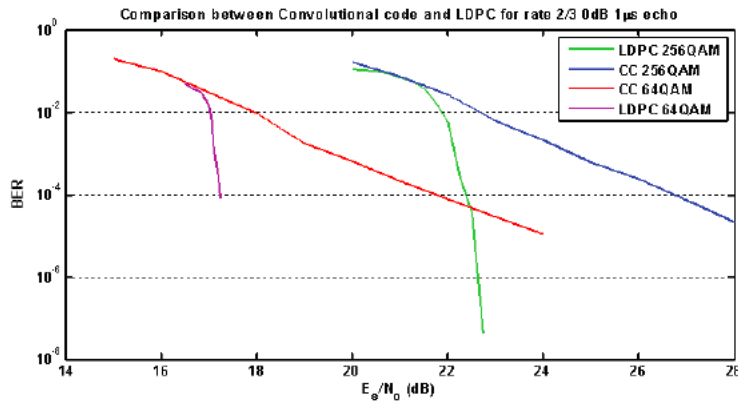


Figure 4.1: BER in function of the  $E_s/N_0$  (actual transmitted power to noise).

Compared with DVB-C, DVB-C2 is no longer restricted to the transmission of MPEG-2 Transport Stream packets but support also other packetized

data streams. Besides the LDPC code, a BCH code is employed by DVB-C2 after LDPC decoding. It adds redundancy to the bit stream which consumes less than 1 percent of the total bit rate. This code with very little error correction capabilities is used to correct the error-floor which typically arises at the output of the LDPC decoder. This error-floor, which occurs in most iterative coding schemes as LDPC or Turbo codes, leads to few remaining bit errors after the decoding process, which cannot be corrected by further iterations of the FEC decoder. The figure 4.1 shows the comparison between the convolutional codes of DVB-T with the LDPC codes of DVB-T2.

In comparison with the first generation digital video broadcasting the significantly increased performance of the forward error correction allows for the application of higher constellation schemes.

In the following paragraphs there is a description for every block of the Bit Interleaved Coding and Modulation, from the Forward Error Correction (FEC), constituted by LDPC and BCH encoding that introduced this section, to the modulation techniques and the Interleaver blocks (Bit Interleaver, Time and Frequency Interleaver).

## 4.1 FEC Encoding

**input** BBFrame made by  $K_{bch}$  bits

**output** FECFrame made by  $N_{ldpc}$  bits

The FEC coding adds at the BBFrames of the input (that contains  $K_{bch}$  bits) a number of:

- $N_{bch}-K_{bch}$  bits with the BCH encoding;
- $N_{ldpc}-K_{ldpc}$  bits with the LDPC encoding.

The resulting codeword always contains either 64800 bits or 16200 bits and is known as a FECFRAME.

In the following lines there are two tables with the FEC coding parameters of the two systems: the first is about normal FECFrame with  $N_{ldpc} = 64800$  bits, the second shows short FECFrame with  $N_{ldpc} = 16200$  bits. These long LDPC codewords improve the performance of the LDPC codes.

In the table of the short FECFrame:

- the code rate is in the column "Effective LDPC Code Rate  $K_{ldpc}/16200$ ";
- the code rate of 4/9 (for the C2-system) and 1/5 (for the T2-system) are only used for protection of L1 pre-signalling and not for data (see chapter 6).

LDPC Code	BCH Uncoded Block $K_{bch}$	BCH coded block $N_{bch}$ LDPC Uncoded Block $K_{ldpc}$	BCH t-error correction	$N_{bch}-K_{bch}$
C2-T2 System				
2/3	43040	43200	10	160
3/4	48408	48600	12	192
4/5	51648	51840	12	192
5/6	53840	54000	10	160
only C2 System				
9/10	58192	58320	8	128
only T2 System				
1/2	32208	32400	12	192
3/5	38688	38880	12	192

Table 4.1: Coding parameters for normal FECFrame with  $N_{ldpc}=64800$  bits.

LDPC Code Identifier	BCH Uncoded Block $K_{bch}$	BCH coded block $N_{bch}$ LDPC Uncoded Block $K_{ldpc}$	BCH t-error correction	$N_{bch}-K_{bch}$	Effective LDPC Rate $K_{ldpc}/16200$
C2-T2 System					
1/2	7032	7200	12	168	4/9
2/3	10362	10800	12	168	2/3
3/4	11712	11880	12	168	11/15
4/5	12432	12600	12	168	7/9
5/6	13152	13320	12	168	37/45
only C2 System					
8/9	14232	14400	12	168	8/9
only T2 System					
1/4	3072	3240	12	168	1/5
3/5	9552	9720	12	168	3/5

Table 4.2: Coding parameters for short FECFrame with  $N_{ldpc}=16200$  bits.

**Outer encoding: BCH****input** BBFrame made by  $K_{bch}$  bits**output** BCHFECFrame made by  $N_{bch}$  bits

The BCH block calculates a new codeword starting from the input bits  $(m_{K_{bch}-1}, m_{K_{bch}-2}, \dots, m_0)$  with the following three processes:

- the polynomial  $m(x) = m_{K_{bch}-1}x^{K_{bch}-1} + m_{K_{bch}-2}x^{K_{bch}-2} + \dots + m_0$  is multiplied for  $x^{N_{bch}-K_{bch}}$  giving a new polynomial  $m'(x)$ ;
- $m'(x)$  is divided for a fixed polynomial  $g(x)$  giving the remainder  $d(x)$ ;
- the coefficients of  $d(x)$  are added to the initial bits of the BBFrame to form the codeword at the output of the BCH block.

**Inner encoding: LDPC****input** BCHFECFrame made by  $N_{bch}$  bits**output** FECFrame made by  $N_{ldpc}$  bits

The process of the Inner encoding is more complex than the Outer encoding and it isn't described (see [3] and [9] for a precise description).

The new bits to be added at the codeword are initialized to 0 and then modified by XOR operations with the bits at the input.

**Bit Interleaver****input** FECFrame made by  $N_{ldpc}$  bits**output** bit stream

The last block of the FEC encoding is the Bit Interleaver. It is divided into two parts: the parity interleaver and the column-twist interleaver. The first operates only on the parity bits, the second writes all the data bits into column that are read-out row-wise.

The Interleaver blocks arrange data in a non-contiguous way to increase performance of the global systems by the reduction to the influence of burst errors or narrow-band interferers.

**4.2 Mapping bits onto constellations****input** bit stream**output** constellations values  $z_q$

After the FEC encoding the bits are demultiplexed into parallel cell words and then mapped into constellation values. The following two tables contain the parameters for the bit-mapping of the two systems.

$\eta_{MOD}$  indicates the effective number of bit per cell.

LDPC block length ( $N_{ldpc}$ )	Modulation mode	$\eta_{MOD}$	Number of output Data Cells
64 800	4096QAM	12	5 400
	1024QAM	10	6 480
	256QAM	8	8 100
	64QAM	6	10 800
	16QAM	4	16 200
16 200	4096QAM	12	1 350
	1024QAM	10	1 620
	256QAM	8	2 025
	64QAM	6	2 700
	16QAM	4	4 050
	QPSK	2	8 100

Table 4.3: C2-System.

As given in the table the modulation schemes of the C2 system varies from 16- to 4096 QAM, while DVB-C maximally employed 256-QAM. The significantly increased performance of the forward error correction allows this application of higher constellation schemes being used for modulation of the OFDM subcarriers (see chapter 6).

LDPC block length ( $N_{ldpc}$ )	Modulation mode	$\eta_{MOD}$	Number of output data cells
64 800	256-QAM	8	8 100
	64-QAM	6	10 800
	16-QAM	4	16 200
	QPSK	2	32 400
16 200	256-QAM	8	2 025
	64-QAM	6	2 700
	16-QAM	4	4 050
	QPSK	2	8 100

Table 4.4: T2-System.

The modulation schemes of the T2 system varies from QPSK to 256 QAM. Compared to DVB-T, DVB-T2 introduces 256-QAM to take full advantage of the efficiency of the error-correction technique. This gives a 33 % increase in spectral efficiency and capacity transported for a given code rate [5].

**Bit to cell word demultiplexer****input** bit stream**output**  $N_{substreams}$  sub-streams in two or more cell words  $(y_{0,q}, \dots, y_{\eta_{mod}-1,q})$ 

The demultiplexer divides the bits into more substreams. The number of the last one depends on the considered modulation and on the  $N_{ldpc}$ .

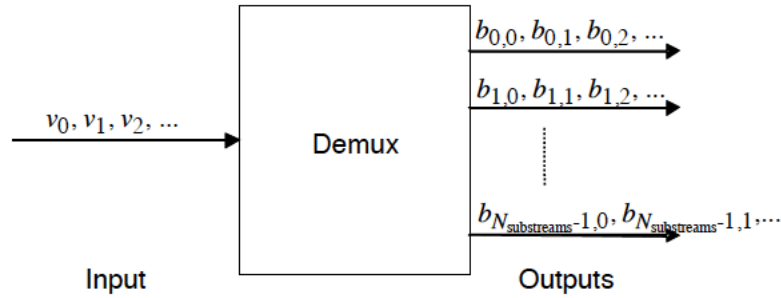


Figure 4.2: Demultiplexer.

$b_{e,do}$  is the output of the demultiplexer,  $e$  is a number defined from  $di$  (the input bit number),  $do = di \text{ div } N_{substreams}$ . An example of the parameters of the C2-system is given in the table 4.5.

Modulation format	64QAM											
Input bit-number, $di \text{ mod } N_{substreams}$	0	1	2	3	4	5	6	7	8	9	10	11
Output bit-number, $e$	11	7	3	10	6	2	9	5	1	8	4	0

Table 4.5: Demultiplexing parameters for code rates  $1/2$ ,  $3/4$ ,  $4/5$ ,  $5/6$ ,  $9/10$ ( $8/9$ ), 64QAM,  $N_{substreams}=12$ .

In the C2-System the words are split into two cell words of width  $\eta_{MOD} = N_{substreams}/2$  (except for the 256QAM with  $N_{ldpc}=16200$  and 4096QAM with  $N_{ldpc}=64800$ ). The first  $\eta_{MOD}$  bits in the first cell, the second  $\eta_{MOD}$  bits in the last cell. For the other two cases the bits are included into cell words of width 8. Only a few configuration are admitted, see the tables 4.6-4.7 (in the T2-system there aren't these restrictions).

**Cell word mapping into I/Q constellations****input**  $N_{substreams}$  sub-streams in two or more cell words  $(y_{0,q}, \dots, y_{\eta_{mod}-1,q})$ **output** constellation points normalized  $f_q$ 

The modulation operates separately for each cell word with constellation:

	Modulation format					
Code rate	QPSK	16QAM	64QAM	256QAM	1024QAM	4096QAM
2/3	NA	NA	X	NA	NA	NA
3/4	NA	NA	NA	X	X	NA
4/5	NA	X	X	NA	NA	NA
5/6	NA	NA	NA	X	X	X
9/10	NA	X	X	X	X	X

Table 4.6: Configurations for  $N_{ldpc}=64800$ .

	Modulation format					
Code rate	QPSK	16QAM	64QAM	256QAM	1024QAM	4096QAM
1/2	N/A	X	NA	NA	NA	NA
2/3	N/A	NA	X	NA	NA	NA
3/4	N/A	NA	NA	X	X	NA
4/5	N/A	X	X	NA	NA	NA
5/6	N/A	NA	NA	X	X	X
8/9	N/A	X	X	X	X	X

Table 4.7: Configurations for  $N_{ldpc}=16200$ .

- QPSK, 16QAM, 64QAM, 256QAM, 1024QAM, 4096QAM for the C2 system;
- QPSK, 16-QAM, 64-QAM, 256-QAM for the T2 system.

The values of  $z_q$ , real and imaginary part, are defined by convention, in the tables 4.8-4.9 there is an example of the 64QAM (C2-System).

$y_{0,q}$	1	1	1	1	0	0	0	0
$y_{2,q}$	0	0	1	1	1	1	0	0
$y_{4,q}$	0	1	1	0	0	1	1	0
$\text{Re}(z_q)$	-7	-5	-3	-1	1	3	5	7

Table 4.8: Constellation mapping for real part of 64QAM.

$y_{1,q}$	1	1	1	1	0	0	0	0
$y_{3,q}$	0	0	1	1	1	1	0	0
$y_{5,q}$	0	1	1	0	0	1	1	0
$\text{Im}(z_q)$	-7	-5	-3	-1	1	3	5	7

Table 4.9: Constellation mapping for imaginary part of 64QAM.

In figure 4.3 an illustration of the 64QAM constellation is given. Each constellation point is normalized to obtain the correct value of  $z_q$ . In the 64QAM the normalized value is  $f_q = \frac{z_q}{\sqrt{42}}$ .

### The additional blocks of the T2-System: Constellation Rotation, Cell/Time Interleaver

Constellation Rotation:

**input** FEC block  $F=(f_0, f_1, \dots, f_{N_{cells}-1})$  ( $N_{cells}$  = number of cells)

**output** output cells  $G=(g_0, g_1, \dots, g_{N_{cells}-1})$

In the T2-system the modulation of the common PLP and of the data PLPs are rotated in the complex plane by a rotation phasor and an assigned angle.

Cell interleaver:

**input**  $G=(g_0, g_1, \dots, g_{N_{cells}-1})$

**output**  $D(r)=(d_{r,0}, d_{r,1}, \dots, d_{r,N_{cells}-1})$

The Cell Interleaver uniformly spreads the cells  $G$  in order to ensure in the receiver an uncorrelated distribution of channel distortions and interference along the FEC codewords. The output of the cell interleaver is:  $d_{r,L_r}(q) = g_{r,q}$  for each  $q = 0, 1, \dots, N_{cells} - 1$ .  $L_r(q) = [L_0(q) + P(r)] \bmod N_{cells}$  is the permutation function where:

- $L_0(q)$  is the basic permutation function and is defined by an algorithm;
- $P(r)$  is a shift value defined by another algorithm;
- $r$  is the number of the FEC block of the TI-block.

The Time Interleaver provides the bulk of the interleaving, spreading the cells of each FEC block over many symbols and potentially over several T2-frames. This offers protection against impulsive interference as well as time-varying channels.

See [9] for the description of this block.



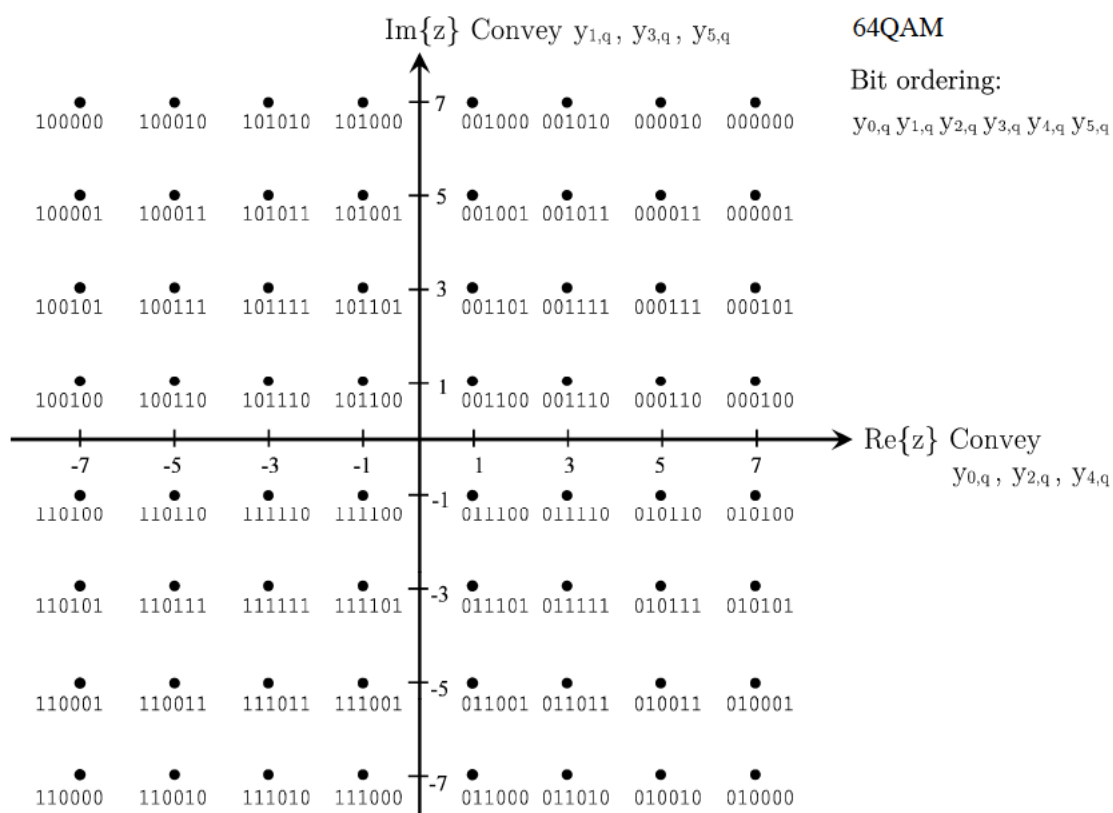


Figure 4.3: 64QAM mapping and the corresponding bit patterns.



## Chapter 5

# Data Slices and L1 signalling

In this chapter there is a description of the C2-system blocks. The corresponding T2-system blocks are considered only in comparison with the C2-blocks.

As previously introduced in Chapter 2, the data slices are groups of PLPs combined together. They are related to channels but their position is no longer coupled to any channel raster: with the use of a shielded environment there is no need to coordinate the spectrum with the external terrestrial environment. However, it's often still present for reasons of compatibility with terrestrial transmission.

Different services are embedded in each Data Slice, that never exceed the bandwidth of an 8 MHz reception tuner. All data required for the reception of one service is transmitted within each Data Slice, that is interleaved in time and frequency for robust protection against interferes or impulsive noise.

A brief description of the Data Slice Packet Generation and of the Data Slice processes is given in the following sections.

### 5.1 Data Slice Packet Generation

**input** cells of the FEC frames

**output** Data Slices

A Data Slice Packet contains the cells of one or two FECFrame. The process of the Data Packet Generation is divided into:

- Data Slice Packets for Data Slice Type 1: transmission of the data packets;
- Data Slice Packets for Data Slice Type 2: synchronization to the Data Slice Packets.

### Data Slice Packets for Data Slice Type 1

**input** constellation points  $f_q$

**output** complex cells  $g_p$

The complex cells at the output are  $g_p = f_q$  with  $q = 0, \dots, N_{DP}-1$  ( $N_{DP}$  is the number of the data cells of one LDPC codeword in a Data Slice Packet). The start of the data transmitted is given by Level 1 Signalling Part 2.

### Data Slice Packets for Data Slice Type 2

A Data Slice Packet Type 2 is given by a FECFrame Header followed by one or two FECFrame. The structure is illustrated in the figure 5.1.

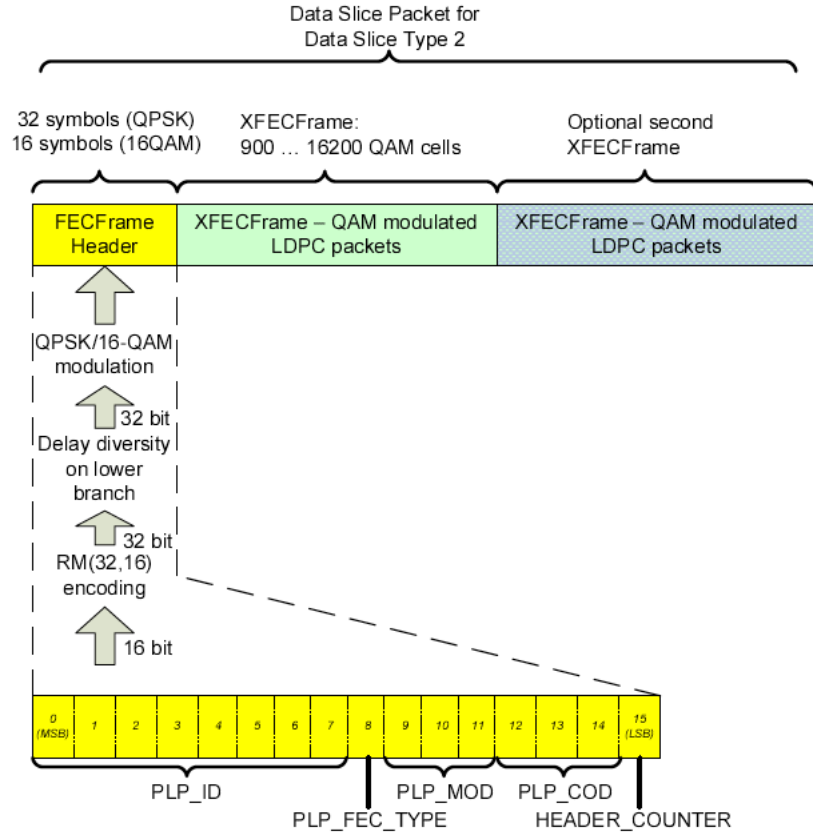


Figure 5.1: Data Slice Packet.

The FECFrame Header contains:

- PLP\_ID (8 bit): the number of PLPs;

- PLP\_FEC\_TYPE (1 bit): the size of the following FECFrame;
- PLP\_MOD (3 bit): the used QAM mappings;
- PLP\_COD (3 bit): the LDPC code rate of the following FECFrame;
- HEADER\_COUNTER (1 bit): the number of FECFrames following this FECFrame header (one or two).

A FECFrame Header is encoded following 3 processes to ensure a robust synchronization and decoding of the L1 signalling part 1 data. The first process is a Reed Muller(32,16) encoding and the resulting 32 bits are subject to two parallel processes. On one side nothing is done. On the other the 32 bit codewords are shifted cyclically and then scrambled. Then the 32 original bits and the simplified/scrambled bits are mapped on 32 symbols QPSK (robust FECFrame header) or 16 symbols 16QAM (high efficiency FECFrame header).

## 5.2 C2-System: Layer 1 part 2 signalling

Figure 5.2 illustrates the C2 Frame structure. The number of the Preamble Symbols depend on the amount of L1 Signalling. L1 signalling part 2 indicates OFDM parameters of the C2 channel and all relevant information for the Data Slices, PLPs and Notch bands.

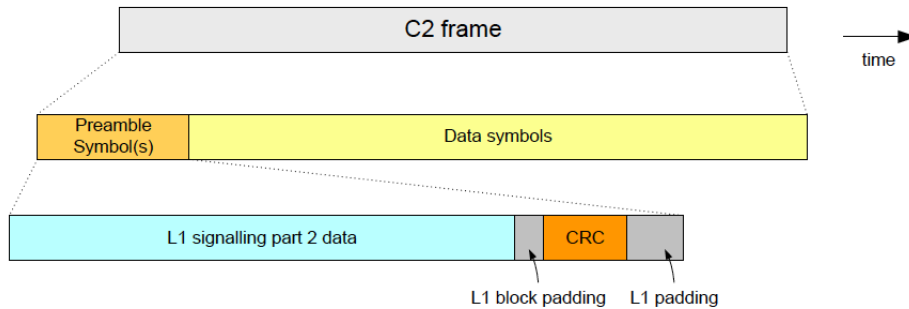


Figure 5.2: L1 part 2 signalling structure.

The preamble header has a fixed length of 32 OFDM cells and is inserted in front of the L1 TI-block. The information bits of the Preamble header are FEC encoded by a Reed-Muller(32,16) code and then encoded by QPSK.

### L1 signalling part 2 data

Figure 5.3 contains the fields for L1 signalling part 2 data.

Field	Size (bits)
NETWORK_ID	16
C2_SYSTEM_ID	16
START_FREQUENCY	24
GUARD_INTERVAL	2
C2_FRAME_LENGTH	10
L1_PART2_CHANGE_COUNTER	8
NUM_BUNDLED_CH	5
NUM_DSLSICE	8
NUM_NOTCH	4
for i=0..NUM_DSLSICE-1 {	
DSLSICE_ID	8
DSLSICE_TUNE_POS	14 or 13
DSLSICE_OFFSET_LEFT	9 or 8
DSLSICE_OFFSET_RIGHT	9 or 8
DSLSICE_TI_DEPTH	2
DSLSICE_TYPE	1
if DSLSICE_TYPE=='1' {	
FEC_HEADER_TYPE	1
}	
DSLSICE_CONST_CONF	1
DSLSICE_LEFT_NOTCH	1
DSLSICE_NUM_PLP	8
for i=0..DSLSICE_NUM_PLP-1 {	
PLP_ID	8
PLP_BUNDLED	1
PLP_TYPE	2
PLP_PAYLOAD_TYPE	5
if PLP_TYPE=='00' or '01' {	
PLP_GROUP_I	8
}	
if DSLSICE_TYPE=='0' {	
PLP_START	14
PLP_FEC_TYPE	1
PLP_MOD	3
PLP_COD	3
}	
PSI/SI_REPROCESSING	1
if PSI/SI_REPROCESSING=='0' {	
transport_stream_id	16
original_network_id	16
}	
}	
RESERVED_1	8
}	
RESERVED_2	8
}	
for i=0..NUM_NOTCH-1 {	
NOTCH_START	14 or 13
NOTCH_WIDTH	9 or 8
RESERVED_3	8
}	
RESERVED_TONE	1
RESERVED_4	16

Figure 5.3: L1 part 2 signalling fields.

The principal fields are described in the list below.

- NETWORK\_ID (16 bit): the current DVB-C2 network;
- C2\_SYSTEM\_ID (16 bit): the C2 System within the DVB-C2 network;
- START\_FREQUENCY (24 bit): the start frequency of the current C2 System (from 0 Hz);

- **GUARD\_INTERVAL** (2 bit): the guard interval of the current C2 Frame;
- **C2\_FRAME\_LENGTH** (10 bit): the number of Data Symbols per C2 Frame;
- **NUM\_DSlice** (8 bit): the number of Data Slices carried within the current C2 Frame.

The following fields appear in the Data Slice loop:

- **DSlice\_ID** (8 bit): it identifies a Data Slice within the C2 System;
- **DSlice\_TYPE** (1 bit): it indicates the type of the associated Data Slice;
- **DSlice\_NUM\_PLP** (8 bit): it indicates the number of PLPs carried within the associated Data Slice;

The following fields appear in the PLP loop:

- **PLP\_ID** (8 bit): it identifies a PLP within the C2 System;
- **PLP\_TYPE** (2 bit): it indicates the type of the associated PLP;
- **PLP\_COD** (3 bit): it indicates the code rate used by the associated PLP.

### **L1 block padding**

This 1-bit field (see figure 5.2) is inserted following the L1 signalling part 2 data to ensure that the length of L1 signalling part 2, given by L1 signalling part 2 data and L1 block padding, is a multiple of 2.

### **CRC for the L1 signalling part 2**

A 32-bit error detection code is applied to the entire L1 signalling part 2 including L1 signalling part 2 data and L1 block padding.

### **L1 padding**

This variable-length field is inserted following the L1 signalling part 2 CRC field to ensure that multiple LDPC blocks of the L1 signalling part 2 have the same information size when the L1 signalling part 2 is segmented into multiple blocks and these blocks are separately encoded.

### Modulation and error correction coding of the L1 part 2 data

Figure 5.4 shows the encoding and modulation of L1 part 2 data. At first a concatenation of a BCH outer code and a LDPC inner code protects the L1 part 2 data. The length of these data bits varies depending on the complexity of the Data Slices considered.

The L1 part 2 data can be segmented into multiple blocks. Since a segmented L1 part 2 data has a length less than BCH information length  $K_{bch} = 7032$ , a shortening operation (zero padding) is required for BCH or LDPC encoding.

After BCH encoding with zero padded information, the BCH parity bits of the L1-part2 data are appended to the L1 part 2 data. The concatenated L1 part 2 data and BCH parity bits are further protected by a shortened and punctured 16K LDPC code with code rate  $1/2$  ( $N_{ldpc} = 16\ 200$ ). Each coded L1 signalling part 2 is bit-interleaved and then is mapped onto constellations.

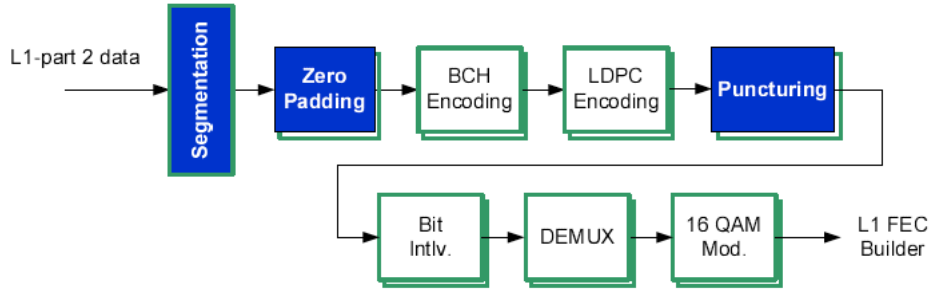


Figure 5.4: Encoding and Modulation of L1 signalling part 2.

Since the length of L1 signalling part 2 is variable, the resulting number of needed L1 frames is also variable. As described in the C2-frame structure (see chapter 6 and figure 6.1), each L1 FECFrame packet corresponds to one L1 block within an OFDM Symbol. As soon as more than one L1 FECFrame packet is necessary, the same number of Preamble Symbols in consecutive OFDM Symbols is needed. If the length of L1 part 2 data exceeds a pre-determined number  $N_{L1part2\_max\_per\_Symbol}$ , the L1 part 2 data shall be divided into equidistant blocks.  $N_{L1part2\_max\_per\_Symbol}$  means the maximum number of L1 information bits for transmitting the coded L1 signalling part 2 through one OFDM Symbol.

### Parameters for FEC encoding of L1 part 2 data

The number of L1 part 2 data bits is variable and the bits are transmitted over one or multiple 16K LDPC blocks depending on the length of the L1 part 2 data. The number of LDPC blocks for the L1 part 2 data,  $N_{L1part2\_FEC\_Block}$ , are determined as follows:



$$N_{L1part2\_FEC\_Block} = \left\lceil \frac{K_{L1part2\_ex\_pad}}{N_{L1part2\_max\_per\_Symbol}} \right\rceil.$$

Where  $K_{L1part2\_ex\_pad}$  denotes the number of information bits of the L1 part 2 signalling excluding the padding field, L1\_PADDING.

$N_{L1part2\_max\_per\_Symbol}$  denotes the maximum number per symbol of the L1 part 2 data.

The length of L1\_PADDING field,  $K_{L1part2\_PADDING}$  is:

$$K_{L1part2\_PADDING} = \left\lceil \frac{K_{L1part2\_ex\_pad}}{N_{L1part2\_FEC\_Block}} \right\rceil \times N_{L1part2\_FEC\_Block} - K_{L1part2\_ex\_pad}. \quad (5.1)$$

The final length of the whole L1 signalling part 2 including the padding field,  $K_{L1part2}$  is:  $K_{L1part2} = K_{L1part2\_ex\_pad} + K_{L1part2\_PADDING}$ .

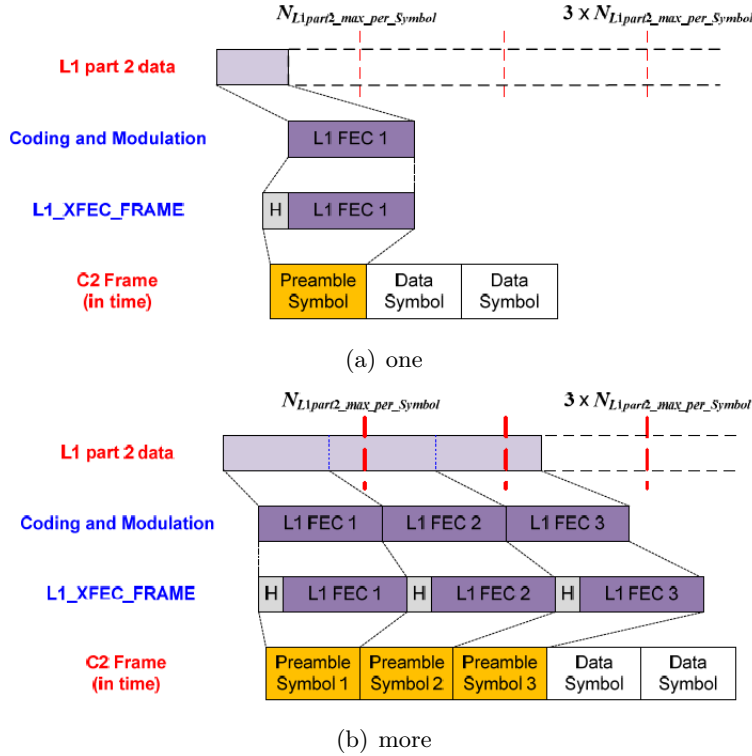


Figure 5.5: a) L1 part 2 fits into one L1 part 2 LDPC FECFrame; b) L1 part 2 fits into more L1 part 2 LDPC FECFrame.

The number of information bits in each of  $N_{L1part2\_FEC\_Block}$  blocks,

$K_{sig}$  is then defined by:

$$K_{sig} := \frac{K_{L1part2}}{N_{L1part2\_FEC\_Block}}.$$

Figure 5.5 shows the process of including the L1 part 2 data into the C2 frame.

### FEC Encoding

**Zero padding of BCH information bits** The  $K_{sig}$  bits are then encoded into a 16K ( $N_{ldpc}=16200$ ) LDPC codeword after BCH encoding.  $K_{sig}$  is always less than the number of BCH information bits ( $=K_{bch}=7032$ ) for a given code rate 1/2, so the BCH code has to be shortened. A part of the information bits of the 16K LDPC code has to be padded with zeros in order to fill  $K_{bch}$  information bits and the padding bits aren't transmitted. A format of data after LDPC encoding of L1 signalling part 2 is given in figure 5.6.

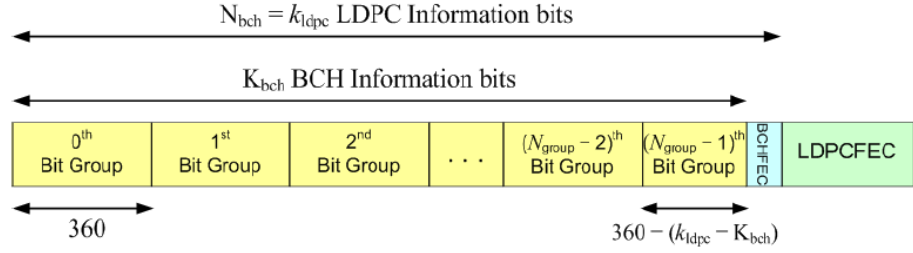


Figure 5.6: Format of data.

The figure shows the BCH information bits divided into groups of 360 bits except for the last group of  $360-(K_{ldpc}-K_{bch})$  bits.

**BCH and LDPC encoding** The  $K_{bch}$  information bits are then BCH encoded to generate  $N_{bch}=K_{ldpc}$  output bits. These bits include the zero padding bits and the BCH parity bits.

**Puncturing of LDPC parity bits** After the LDPC encoding some of the LDPC parity bits has to be punctured: these bits aren't transmitted. In the figure 5.7 is given a representation of the parity bits groups in a FEC block.

The figure shows that the parity bits are divided into  $Q_{ldpc}$  parity groups, each formed by 360 bits.

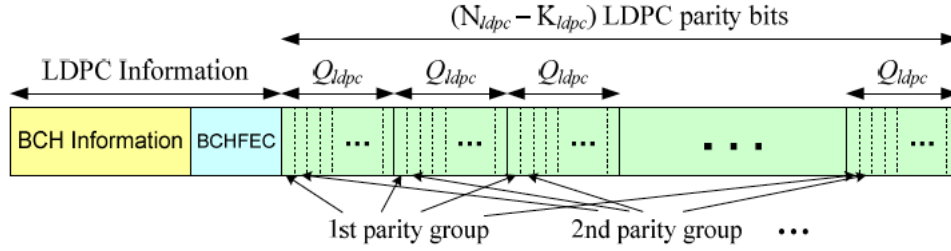


Figure 5.7: Format of data.

**Removal and Interleaving** The zero padding bits are removed and hasn't to be transmitted. At the end the bits are interleaved.

### Mapping bits onto constellations

The bit-interleaved LDPC codewords are then mapped into constellation. The L1 signalling part 2 is first demultiplexed into cell words and then these words are mapped into constellations. For the details of these two operations see [3].

### Time interleaving of L1 signalling part 2 data

The Time Interleaver provides protection of the signal against impulsive interference as well as time-varying channels. L1-part2 data transmission is more robust than the payload data, especially when the time interleaving is applied to the Data Slice.

## 5.3 T2 system: L1 signalling

In the T2 system the L1 signalling provides the receiver with a means to access physical layer pipes within the T2-frames. The L1 signalling structure is split into three main sections: the P1 signalling, the L1-pre signalling and L1-post signalling.

The purpose of the P1 signalling, which is carried by the P1 symbol, is to indicate the transmission type and basic transmission parameters. The remaining signalling is carried by the P2 symbol(s), which may also carry data. The information given by the P1 signalling are divided into 2 types: the first type distinguishes the preamble format, the second type helps the receiver to rapidly characterize the basic transmission parameters.

The L1-pre signalling enables the reception and decoding of the L1-post signalling, which in turn conveys the parameters needed by the receiver to access the physical layer pipes.

The L1-post signalling gives the sufficient information to the receiver to decode the desired PLP.

## Chapter 6

# Frame Builder and OFDM generation

### 6.1 C2 vs T2 Frame structure

**input** Data Slices, L1 header and data (C2 system) - PLPs and L1 signalling (T2 system)

**output** Frame

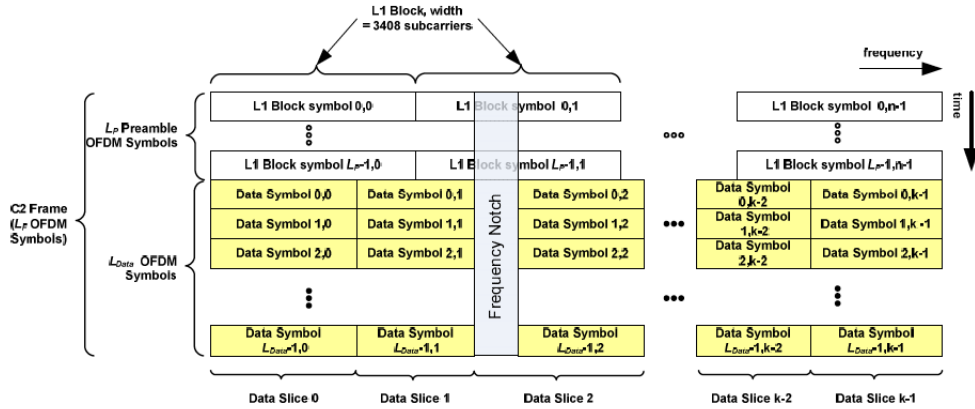


Figure 6.1: C2 Frame Structure.

The Frame Builder assembles the cells of data slices (PLPs in the T2 system) and L1 signalling into arrays corresponding to OFDM symbols.

The C2 frame structure is given in figure 6.1. In the frequency direction there are  $L_p$  Preamble OFDM Symbols and  $L_{data}$  OFDM Symbols, all together form the C2 Frame. In the time direction there are data slices and frequency notches. The transmitter repeats the L1 blocks cyclically every

7.61 MHz giving the possibility to access the complete L1 part 2 signalling in any tuning position of an 8 MHz receiver tuner. As given in figure 6.2 the receiver is able to restore the complete data of a L1 Block by re-ordering the OFDM carriers after converting them into the frequency domain (see the OFDM section). Even the loss of some carriers would not seriously affect the system performance, as the signalling data is transmitted in a very robust mode.

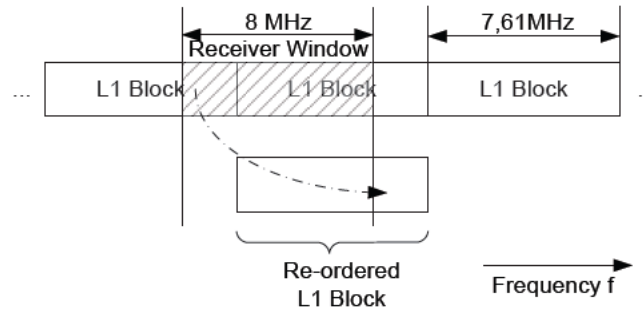


Figure 6.2: Concept of the frequency cyclic L1 signalling and the reordering of the L1 Blocks to obtain the complete L1 signalling.

This structure includes cable specific optimizations.

The Preamble carries the L1 signalling, required for the demodulation of the payload data. L1 signalling gives an improved flexibility: the frequency and the bandwidth of the Data Slices may vary between different C2 frames. It also contains the start and the end frequency of the Data Slices and the optimal tuning position. Each Data Slice and C2 frame is dynamically adaptable to the specific needs without any required re-tuning of the receiver.

The T2 Frame structure is divided, from the highest to the lowest level, into Super Frames, Frames, P1,P2 symbols and Data symbols.

The T2 frame structure is given in figure 6.3.

The T2 Super Frames contains T2-Frames and FEF parts. The L1 pre-signalling indicates the number of Frames in a Super Frame; the L1 post-signalling indicates the current Frame (see paragraph 5.3). The T2-frame comprises one P1 preamble symbol, followed by one or more P2 preamble symbols, followed by a configurable number of data symbols.

While in the C2 system the L1 blocks are responsible of the signalling (by prepending a preamble with the L1 signalling information as seen in the frame structure), in the T2-system the L1 signalling is carried by the P2 symbols, the time and the frequency synchronization are given by the P1 and the P2 symbols. All the estimation process is given by the P2 symbol. The P1 symbol has a particular structure designed to improve the robustness of the P1 detection in the presence of the most challenging channels with

opposite phase; the detection of the P1 symbol is also used to derive the initial time and frequency reference.

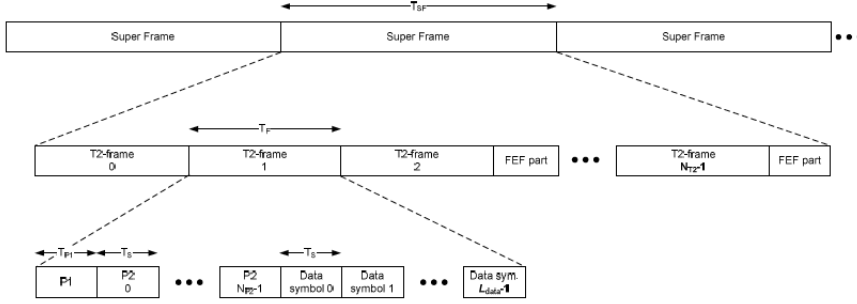


Figure 6.3: T2 Frame Structure.

## 6.2 OFDM generation

The main feature of the DVB-x2 is the application of OFDM. The huge variety of services, like the "Triple Play" packages (telephony, Internet Access, TV) and the HD, requires a downlink spectrum larger than the actual one. The most promising solution to face this demand is the application of a more spectrum efficient physical layer. Following this way, T2 provides the application of 1024 and 4096 QAM in addition to a powerful coding scheme based on LDPC codes as explained in chapter 4.

In this section a description of another powerful solution for the transmission on dispersive channels is given: the OFDM. Its robustness to different types of channel impairments (see also section 7.1) and its flexibility in providing large number of services make this one of the most powerful instruments of the second generation DVB.

### Definition of OFDM

Orthogonal frequency-division multiplexing (OFDM) is a frequency - division multiplexing (FDM) scheme used as a digital multi-carrier modulation method. A large number of closely-spaced orthogonal sub-carriers are used to carry data. The data is divided into several parallel data streams or channels (see figure 6.4), one for each sub-carrier.

The use of these subcarriers gives a solution to the problem that in a wideband channel the gain  $g_{ch}(t)$  is dispersive and causes intersymbol interference (ISI). The channel frequency response shows selectivity and some frequencies suffer deeper attenuation than others. The solution to this problem is to divide the frequency domain in small frequency bands of width  $B_0 = 1/T_0$  (with  $T_0$  the symbol period), at a center frequency of 0 MHz

(DVB-C2) or a generic frequency  $f_i$  (DVB-T2). This partition is so small to assure a quasi-constant gain  $C_i \simeq G_{ch}(f_i)$  for each subband.

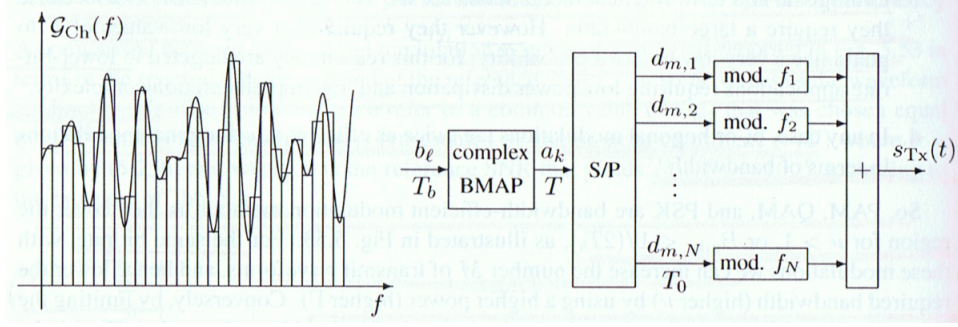


Figure 6.4: The subcarriers in the OFDM.

The bit sequence  $b_l$  that enters in the OFDM generation is first mapped into a complex valued symbol sequence  $a_k = a_{k,I} + ja_{k,Q}$ , typically with the QAM constellation. Then the S/P serial to parallel block separates the symbols in the frequency bands. Every subcarrier (or frequency band) has a modulator that operates independently from the other modulators. At the end all the data are added to form the signal that is transmitted on the channel.

### DVB-T2 and MISO processing

In DVB-T2 a MISO system can be applied in the OFDM generation, that allows simultaneous transmissions on the same frequency avoiding destructive interference. In this configuration the data on different transmitters are not identical but closely related and the pilots are arranged much than at the receiver. An estimate of each transmit channel can be obtained. The idea of this process is given in figure 6.5.

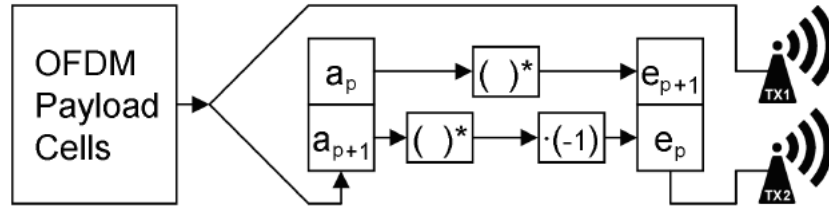


Figure 6.5: MISO processing of OFDM payload cells.

The MISO processing is applied on the cell level. All DVB-T2 receivers have to be able to receive signals with MISO processing. The input data cells are processed, divided into two sets and then directed to two groups



of transmitters. The encoding process is done on pairs of OFDM payload cells  $(a_{m,l,p}, a_{m,l,p+1})$ . The encoded OFDM payload cells  $e_{m,l,p}(T \times 1)$  for MISO transmitter group 1 and  $e_{m,l,p}(T \times 2)$  for MISO transmitter group 2 are generated from the input cells in this manner:

$$\begin{aligned} e_{m,1,p}(T \times 1) &= a_{m,1,p} & e_{m,1,p+1}(T \times 1) &= a_{m,1,p+1} \\ e_{m,1,p}(T \times 2) &= -a_{m,1,p+1}^* & e_{m,1,p+1}(T \times 2) &= a_{m,1,p}^* \end{aligned}$$

with:  $p = \{0, 2, 4, 5, \dots, N_{data} - 2\}$

where  $*$  denotes the complex conjugation operation and  $N_{data}$  is the number of cells at the frequency interleaver output (the previous block in the DVB-T2 system).

The transmitters of the first group don't change the input cells.

MISO processing isn't applied to the P1 symbol. The contents of the P1 symbol will be identical between the two groups of transmitters.

When MISO is not used, the input cells are copied directly to the output, i.e.  $e_{m,l,p} = a_{m,l,p}$  for  $p = 0, 1, 2, \dots, N_{data}-1$ .

### Scattered and Continual Pilots

The Pilot insertion is one of the principal blocks of the OFDM generation in C2/T2-system. The pilot patterns (scattered and continual) are important at the receiver for correctly detect the transmitted information.

The scattered pilot (SP) sequences modulate a set of equally spaced subcarriers providing a reliable channel estimate. They are inserted into the signal at regular intervals in both time and frequency directions and are used by the receiver to estimate changes in channel response in the same two dimensions.

In the T2-system the pilots give the frame, frequency and time synchronization, the channel estimation, the transmission mode identification and are also used to follow the phase noise. Figure 6.6 shows the presence of the various types of pilots in each type of symbol.

Symbol	PILOT TYPE				
	Scattered	Continual	Edge	P2	FRAME-CLOSING
P1					
P2				X	
Normal	X	X	X		
Frame closing			X		X

Figure 6.6: Different types of pilot and symbols in which they appear.

The reference information is given by the reference sequence (see [9] for the definition) and is transmitted in scattered pilot cells in every symbol

on the frame, except P1, P2 and the closing symbol (if applicable) of the T2-frame. The symbol  $l$  is a scattered pilot if satisfies:

$$\begin{aligned} k \bmod (D_X D_Y) &= D_X(l \bmod D_Y) \text{ [normal carrier mode]} \\ (k - K_{ext}) \bmod (D_X D_Y) &= D_X(l \bmod D_Y) \text{ [extended carrier mode]} \end{aligned}$$

where:

- $l$  = denotes the OFDM Symbol number starting from 0 for the first Preamble Symbol of the frame in the C2 system, from 0 for the first P2 symbol of the frame in the T2 system;
- $k$  denotes the carrier number;
- $D_X$  is the separation of pilot bearing carriers (see table 51, [9]);
- $D_Y$  is the number of symbols forming one scattered pilot sequence (see table 51, [9]);
- $K_{ext}$  is the number of carriers added on each side in extended carrier mode (see table 60, [9]);
- $k \in [K_{min}, K_{max}]$  with  $K_{min}$  and  $K_{max}$  are respectively the minimum and the maximum subcarriers of the OFDM generation.

The DVB-C2 uses the same scattered pilot patterns, which allow a common channel estimation block for both systems.

The continual pilots provide a means for fine frequency synchronization and common phase error correction.

The locations of the continual pilots are taken from one or more "CP groups" depending on the FFT mode, see [9] for a precise description.

## IFFT-OFDM Modulation

This paragraph gives a detailed description of the OFDM modulation. The transmitted signal is organized in frames. Each frame has duration of  $T_F$ , and consists of  $L_F$  OFDM Symbols. In the T2-system  $N_{T2}$  frames constitute one superframe.

Each symbol is constituted by a set of  $K_{total}$  carriers transmitted with a duration  $T_S$ . It is composed of two parts: a useful part with duration  $T_U$  followed by a guard interval with duration  $\Delta$ . The symbols in a C2/T2 Frame are numbered from 0 to  $L_F-1$  (excluding P1 symbol in the T2 system). All symbols contain data and reference information. Since the OFDM signal comprises many separately modulated carriers, as explained in the previous paragraph, each symbol can in turn be considered to be divided into cells, each corresponding to the modulation carried on one carrier during one symbol. The carriers are indexed by  $k \in [K_{min}; K_{max}]$  and determined by  $K_{min}$  and  $K_{max}$ . The spacing between adjacent carriers is  $1/T_U$  while the spacing

between carriers  $K_{min}$  and  $K_{max}$  are determined by  $K_{total}/T_U$  in the C2 system and by  $(K_{total}-1)/T_U$  in the T2 system.

The transmitted signal has the following expressions:

$$s(t) = \Re \left\{ \sum_{m=0}^{\infty} \left[ \frac{1}{\sqrt{K_{total}}} \sum_{l=0}^{L_F-1} \sum_{k=K_{min}}^{K_{max}} c_{m,l,k} \times \Psi_{m,l,k}(t) \right] \right\} \quad (6.1)$$

where:

$$\Psi_{m,l,k}(t) = \begin{cases} e^{j2\pi \frac{k}{T_U}(t-\Delta-lT_S-mT_F)} & mT_F+lT_S \geq t \geq mT_F+(l+1)T_S \\ 0 & \text{otherwise} \end{cases} \quad (6.2)$$

These two equations relate to the C2 system. For the T2 system:

$$s(t) = \Re \left\{ e^{j2\pi f_c t} \sum_{m=0}^{\infty} \left[ p_1(t - mT_F) + \frac{5}{\sqrt{27 \times K_{total}}} \sum_{l=0}^{L_F-1} \sum_{k=K_{min}}^{K_{max}} c_{m,l,k} \times \Psi_{m,l,k}(t) \right] \right\} \quad (6.3)$$

Where:

$$\Psi_{m,l,k}(t) = \begin{cases} e^{j2\pi \frac{k'}{T_U}(t-\Delta-T_{P1}-lT_S-mT_F)} & [*] \\ 0 & \text{otherwise} \end{cases} \quad (6.4)$$

$$[*] = mT_F+T_{P1}+lT_S \geq t \geq mT_F+T_{P1}+(l+1)T_S$$

- $l, k$  as defined in the previous paragraph;
- $m$  denotes the C2/T2-frame number;
- $K_{total}$  is the number of transmitted carriers;
- $L_F$  number of OFDM symbols per frame;
- $T_S$  is the total symbol duration for all symbols (except P1 in the T2-system), and  $T_S = T_U + \Delta$ ;
- $T_U$  is the active symbol duration;
- $\Delta$  is the duration of the guard interval;
- $c_{m,l,k}$  is the complex modulation value for carrier  $k$  of the OFDM Symbol number  $l$  in C2/T2 Frame number  $m$ ;

- $T_F$  is the duration of a frame,  $T_F = L_F T_S$  for the C2 system,  $T_F = L_F T_S + T_{P1}$  for the T2 system;
- $K_{min}$  carrier index of first (lowest frequency) active carrier;
- $K_{max}$  carrier index of last (highest frequency) active carrier.

For the T2 system there are parameters related to the P1 symbol and to the center frequency that isn't zero like in the C2 system:

- $f_c$  is the central frequency of the RF signal;
- $k'$  is the carrier index relative to the centre frequency;
- $T_{P1}$  is the duration of the P1 symbol;
- $p_1(t)$  is the P1 waveform.

For all the OFDM parameters see [3] and [9].

### Peak to Average Power Ration (PAPR) reduction

Let  $s(t)$  be a baseband digitally modulated signal. The PAPR of the signal  $s(t)$  is defined as the ration between the maximum power and the average power of the signal:

$$PAPR\{s(t)\} = \frac{\max |s(t)|^2}{E\{|s(t)|^2\}}$$

Further, when the number of sub-carriers increases, the PAPR also increases. For example in a OFDM system with  $N$  subcarriers the maximum power occurs when all the subcarriers with identical phases are added together,  $\max |s(t)|^2 = N$ . When  $E\{|s(t)|^2\} = 1$  the maximum power is  $N$  times the average power of the signal ( $PAPR = N$ ), as  $N$  increases the maximum power becomes larger. Since both the transmitter and receivers amplifiers operate at best when they are near the saturation, having large PAPR means that most of the time the amplifiers are not efficient.

So the sum of many subcarrier components with the IFFT operation leads to high peak values of the transmit signals in the time domain. For this reason the problem of PAPR reduction becomes important with the second generation DVB that uses OFDM in place of single carrier modulation.

The principal techniques of the PAPR reduction are:

- clipping technique: clipping of the signal around the peaks;
- coding technique: selection of the codewords that minimize or reduce the PAPR;

- probabilistic (scrambling) technique: scrambling an input data block of OFDM symbols and transmitting one of them with the minimum PAPR so that the probability of incurring high PAPR can be reduced;
- adaptative predistortion technique;
- DFT-spreading technique: spreading the input signal with DFT (Discrete Fourier Transform).

One simplest approach of reducing the PAPR is to clip the amplitude of the signal to a fixed level  $\mu$ . In other words, any signal whose amplitude exceeds  $\mu$  will saturate its amplitude to the clipping level  $\mu$ . A lower clipping level  $\mu$  gives a good PAPR but also an high clipping distortion of the signal. It also may cause in-band and out-of-band interferences while destroying the orthogonality among the subcarriers.

The coding technique causes no distortion and creates no out-of-band radiation, on the other side it suffers to find the best codes and to store large lookup tables for encoding and decoding, especially for a large number of subcarriers.

The scrambling technique does not suffer from the out-of-band power. However, the spectral efficiency decreases and the complexity increases as the number of subcarriers increases. The main limitation of this technique is that it cannot guarantee the PAPR below a specified level.

The principal techniques for PAPR reduction of the T2-system are the reserved - carrier PAPR reduction and the active constellation extension (ACE). Both of them belong to the class of adaptive predistortion technique. Reserved - carrier PAPR reduction sacrifices a small amount of throughput by reserving some subcarriers which do not carry data. The Active constellation extension modifies some of the transmitted constellations by selectively moving their outer points to positions having greater amplitude.

The dummy carrier reservation, done at the Frame Builder level, is used for PAPR reduction in C2 and is similar to the reserved-carrier PAPR reduction of the T2-system. Some OFDM cells are reserved: figure 6.7 shows a possible implementation for PAPR reduction of the C2-system.

Firstly reserved carriers are allocated. As in the T2-system these reserved carriers don't carry any data information and are instead filled with a peak-reduction signal in order to reduce PAPR. After the IFFT, peak cancellation is operated to reduce PAPR by using a predetermined signal. The predetermined signal, or kernel, is generated by the reserved carriers.

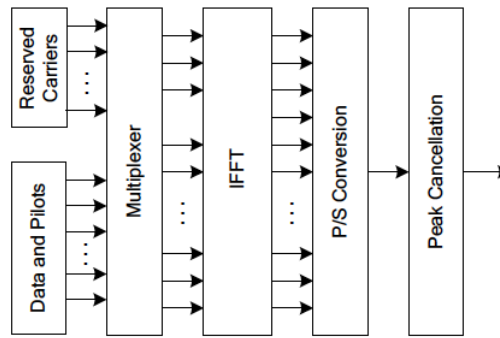


Figure 6.7: The structure of the OFDM transmitter with PAPR reduction using reserved carriers.

## Chapter 7

# Channel Model

The channel is used to convey an information signal from one or more transmitters to one or more receivers. It can be a physical transmission medium such as a wire or a logical connection over a multiplexed medium such as a radio channel. A channel has a certain capacity for transmitting information measured by its bandwidth in Hz or its data rate in bits per second.

All the variables and error characteristics of a channel can be represented in a theoretical channel model. In the following sections the two channel models are described: the HFC and the wireless channel model.

### 7.1 HFC Channel Model

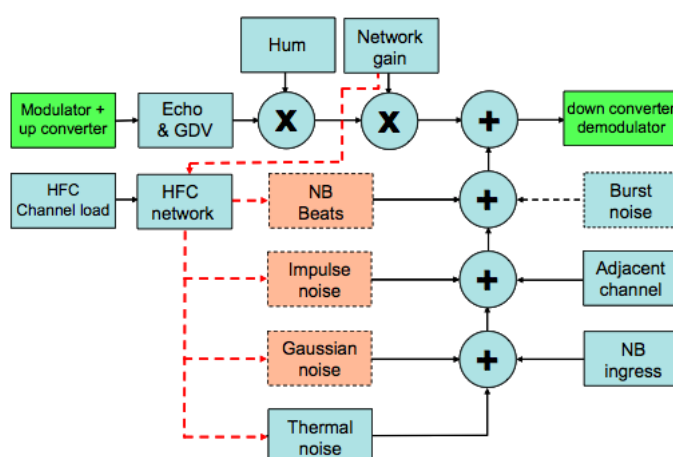


Figure 7.1: Block diagram of the HFC channel model.

A block diagram of the HFC Channel Model is given in the figure 7.1. The channel model is composed of a basic model in combination with additional components. The next paragraphs give the description of some blocks.

## Echo

As specified in [7], the cable channel is modeled as a network of coaxial cables connecting components. These connectors may cause reflections of the signal, i.e. signals traveling in the opposite direction of the transmitted signal. The echo is a particular linear distortion of the signals during its transport from the transmitter to the receiver that describes this phenomenon. It's caused by impedance transitions occurring at connectors, components, damaged cables, incorrectly (or not) terminated cables. For example, reflections with a delay of 100 ns and longer appear in the video image as a horizontal blurring and sometimes a secondary image can be seen. Reflections with a delay shorter than 50 ns do not contribute to the blurring, only short delays may affect the color of the image.

Part of the signal power reflected can rebound in the correct direction and reach the receiver in different intervals of time. Furthermore, the parts of signal reflected twice have a signal loss equal to twice a reflection loss plus the attenuation of passing the coaxial cable twice. The reception at different times of attenuated signal parts causes a difficult detection of the signal at the receiver.

Length segment [m]	250		50		10		2	
Frequency [MHz]	140	800	140	800	140	800	140	800
Reflection loss (2x) [dB]	36	20	36	20	36	20	36	20
Attenuation @ 3dB/100m (200MHz) [dB]	15	30	3	6	0,6	1,2	-	-
Total Loss [dB]	51	50	39	26	37	21	36	20
Delay [ns]	2500		500		100		20	

Figure 7.2: Attenuation and delay of forward reflections.

In the figure 7.2 the data shows that for length segment of 250 m the total loss is similar for the two given frequencies. This is the case of the HFC network, that is composed of long segments interconnecting the amplifiers, so that reflections do not contribute to the echo. However in the case of short coaxial connections of 2 m and less, the reflections significantly contribute to the echo. Since the range between 10 m and 50 m includes segments that can be found in the in-home network, it's important to considerate the reflections' consequences to the echo in this case.

In the home network the major source for signal reflections is the use of components like resistive and inductive splitters. A typical case of the home



networks is given in figure 7.3 for an inductive splitter.

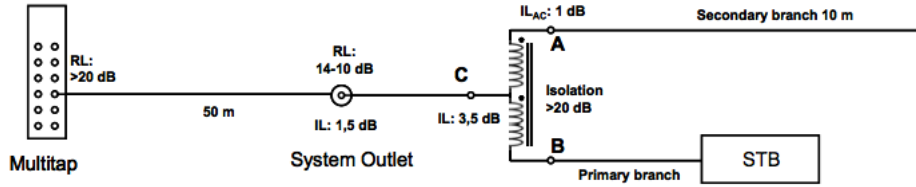


Figure 7.3: Typical electric scenario for an inductive splitter.

This electric scenario can present three situations:

- Part of the reflected signal at point A will be inserted in the primary cable at point B because of the limited isolation of both ports. In the worst case the full signal power inserted in the secondary branch connected to port A will be reflected: the signal with a signal power reduced is injected in the primary branch.
- Part of the reflected signal at port A will be passed on to the input port C toward the system outlet. A fraction of this signal is then reflected at the TV out connector and conveyed to port B through the splitter.
- The reflected signal from the secondary branch will pass with some loss the system outlet and is reflected at the multi-tap.

The values that characterize this system are the loss through the path from port A to port B, the reflection via system outlet and the reflection via multi-tap.

In the measurements of the Redesign Project echoes with power levels of up to 6 dB below the main signal have been found. However, as the typical in-home cabling is limited to few meters only, the delay of the echoes is far less than  $1\mu\text{s}$ .

The use of OFDM improved the robustness of the cable channel against echoes.

## Gaussian noise

The Gaussian noise consists of thermal noise and intermodulation products.

The intermodulation products are distortion signals associated with the nonlinear behavior of components. Amplifiers, optical transmitters and receivers are non-linear components, their transfer function can be expressed as a mathematical expansion:

$$y(t) = a_0 + a_1x(t) + a_2x^2(t) + a_3x^3(t) + a_4x^4(t) + a_5x^5(t) + a_6x^6(t) + \dots \quad (7.1)$$

with:

- $y(t)$  the momentary output level
- $x(t)$  the momentary input level
- $a_i$  parameters

The signal power of the DVB-C signal is approximately equally distributed over the whole frequency band. Since in the frequency domain the PSD of the product is given by the convolution of the PSD's of the parents, the IM product with one or more digital parents has a relatively "broad" spectrum with a low PSD. In most HFC frequency plans the signals are spaced following this equation:  $n \times f_{spacing} + \delta$  with  $f_{spacing} = 8\text{MHz}$ . The IM products of all signals are so situated at a limited number of well defined frequencies with an offset of  $m \times \delta$  ( $m = 0, 1, 2, \dots$ ) with reference to the carrier frequency of the signal.

For IM products with a broadband parent the signal will be distributed over a frequency band much larger than  $\delta$ . In the time domain DVB-C signals do not have a periodic shape but they can be considered more noise-like as compared to PAL video.

From the HFC network viewpoint, the ratio of the DVB-C2 signal power level ( $P_{DVB-C2}$ ) and of the power level of the composite 2<sup>nd</sup> and 3<sup>rd</sup> order IM products in a channel ( $P_{IM2}$ ,  $P_{IM3}$ ) has to be considered as a network engineering parameter. The operator has to understand their nature in the frequency and time domain, and to find the appropriate balance between network load and the  $P_{IM2}$ ,  $P_{IM3}$  distortion levels.

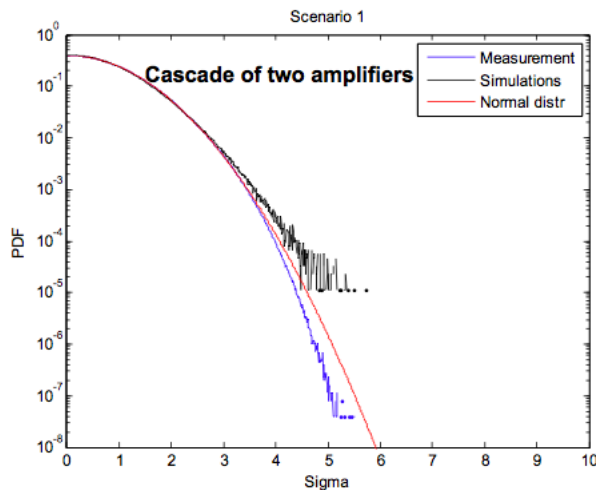


Figure 7.4: Signal level probability density of intermodulation products for a cascade with two amplifiers and loaded with a full package of analogue and digital services.

The Telenet lab (see [7]) gave measurements and simulations of the 2<sup>nd</sup> and 3<sup>rd</sup> order IM distortion signals for the cascade of two amplifiers. The PDF were calculated and plotted in the figure 7.4. Both the measured and simulated data perfectly agree with the normal distribution up to a signal level of 4 times the RMS noise level. This consideration is valid also for the cases of an optical link and an optical link with an amplifier: the 2<sup>nd</sup> and 3<sup>rd</sup> order IM products are additive white Gaussian noise (AWGN) like in nature.

The OFDM has improved cable network transmissions especially in presence of non - linearities of the active network components, as the Gaussian noise. One coaxial cable transmits multiple signals, the sum of the signals forms a probability density function close to the Gaussian distribution. So the resulting disturbances in full digital networks are like white Gaussian noise, which effects single-carrier or multi-carrier signals in the same way. Measurements indicate that the disturbs are mainly narrow-band, so that only a few sub-carriers will be affected and not the complete single-carrier signal as it was the case for the transmissions of the first generation digital video broadcasting.

### Impulsive noise

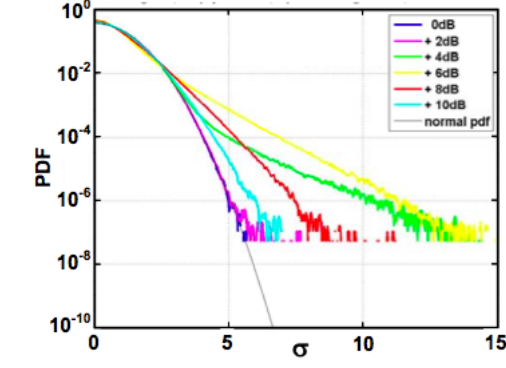
An impulse noise can be defined as a single sequence of events with short duration that occur randomly distributed in the time domain. The measurements given in [7, pp. 23-24] show that impulse events generated by external sources are rather limited: in the case of good HFC network, these events are not a major source for performance degradation.

The Kathrein studies (see [7]) show the generation of impulse events in case of overloading the components. Measurements for an optical link with full digital load for increasing output power levels are given in Figure 8.5 a), where we show the PDFs of the distortion signals as a function of the RMS noise level. This output signal level is measured relative to the lowest output level for which the studies have performed a measurement. All PDFs have been normalized to the average RMS level of the captured distortion signal. The data substantially agree with the normal distribution, except for the range from +4 up to +8 dB: this deviation demonstrates the occurrence of impulse-like events.

Other measurements give the idea of the decomposition of  $P_{IM}$  into  $P_{gaussian}$  and  $P_{impulse}$ , where  $P_{IM}$  is the signal power associated with the IM products,  $P_{gaussian}$  with the additive Gaussian-distribution and  $P_{impulse}$  with the impulse events. Figure 7.5 b) shows that only a small, or a minor fraction at most, can be attributed to impulse events whereas the major part is associated with intermodulation products with a normal distribution.

The analysis of the cascade of 9 amplifiers shows that the deviation is much less pronounced respect of the optical link. In this case considering

that for a signal load of +8 dB or less the deviation is insignificant and that the maximum SNR of the cascade was found for a channel load of +4dB, a cascade of amplifiers that works correctly does not yield impulse-like distortion events.



(a) PDF distortion signals

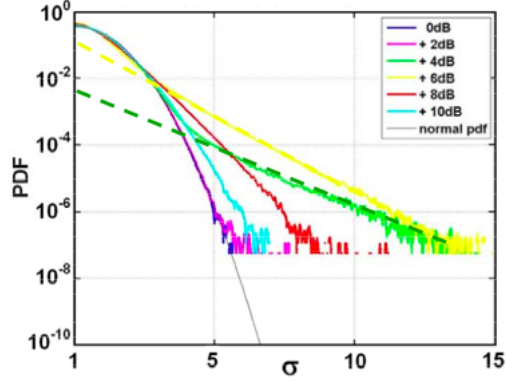
(b)  $P_{IM}$  into  $P_{gaussian}$  and  $P_{impulse}$ 

Figure 7.5: a) PDF of the distortion signals produced by an optical link loaded with 96256 QAM channels; b) Decomposition of  $P_{IM}$  into  $P_{gaussian}$  and  $P_{impulse}$ .

The studies of the distortion products generated by an in-home amplifier with an all digital and a mixed digital analogue load show a strong deviation from the normal distribution. The impulse noise is still present. Other studies on the distortion signal generated in the customer home domain show that switching on/off of electronic equipment generates impulse events.

OFDM collects all impulsive noise events during the OFDM symbol duration. In the frequency domain at the receiver side the collected noise is almost equally distributed over all OFDM sub-carriers: strong impulsive noise can result in loss of the complete OFDM symbol. As described in the

first part of this paragraph the impulsive events are rather limited and very short, so the resulting disturbance is less compared to other forms of noise as the thermal noise.

### **Narrow band beats**

Narrow band beats may occur only above the regular working point of the channel load. These events can be avoided by an appropriate design of the signal levels along the cascade.

### **Burst noise**

Burst events can be define as single events with long duration that occur randomly distributed in the time domain. Typically they are formed by a large number or "train" of impulses within a short period (the burst), followed by long periods without any activity. The measurements given in [7, pp. 23-24] show that burst events were not observed. Measurements on the distortion signals generated in the customer home domain show that switching on/off electronic in the vicinity of the coaxial cable feeding a DVB-C receiver generates burst events.

## **7.2 Wireless Channel**

For a good deriving of a wireless channel model, it's important to distinguish the distortions that characterize the channel. It is no simple to analyze all the parameters that influence the channel, so the main goal of this kind studies relating the wireless channel is to determine the most significant parameters that give an approximated model. There are two types of variations of the channel strength over time and over frequency:

- Large-scale fading, due to the path loss of signal as a function of distance and shadowing by large objects such as buildings and hills. An example can be mobile phones that move through a distance of the order of the cell size.
- Small-scale fading, due to the constructive and destructive interference of the multiple signal paths between the transmitter and receiver. This occurs at the spatial scale of the order of the carrier wavelength.

Large-scale fading is frequency independent and more relevant to issues such as cell-site planning; small-scale fading is frequency dependent and more relevant to the design of reliable and efficient communication systems.

The first part of this section is dedicated to physical modeling, the second part is dedicated to the derivation of an input-output wireless channel model. The main difference between the HFC and the wireless channel model is that

the first is time-invariant over a very long time-scale, while wireless channels are typically time-varying. The models depend very much on the time-scales of interest, when it's possible the time-variant situations considered will be approximated into time-invariant ones.

### Physical modeling for wireless channels

A wireless signal is associated to electromagnetic waves, that propagate from the transmitter to the receiver. The correct transmission of the signal depends on a lot of parameters that vary by the situation considered. For a first example consider a fixed antenna radiating into free space. In a field sufficiently far away from the antenna the electric field, the magnetic field and the direction of propagation from the antenna are perpendicular all together. They are also proportional to each other. In response to a transmitted sinusoid  $\cos 2\pi ft$ , the expression of the electric field at time  $t$  is:

$$E(f, t, (r, \theta, \psi)) = \frac{\alpha_s(\theta, \psi, f) \cos 2\pi f(t - r/c)}{r}. \quad (7.2)$$

where:

- $u=(r, \theta, \psi)$  is the point in space at which the electric field is being measured;
- $r$  is the distance from the transmit antenna to  $u$ ;
- $(\theta, \psi)$  is the vertical and horizontal angles from the antenna to  $u$  respectively;
- $c$  is the speed of light;
- $\alpha_s(\theta, \psi, f)$  is the radiation pattern of the sending antenna at frequency  $f$  in the direction  $(\theta, \psi)$ .

There are two points that arise from this expression: the phase of the field that varies with  $fr/c$  and the fact that as the distance  $r$  increases the electric field decreases as  $r^{-1}$ . The first indicates the delay caused by the radiation traveling at the speed of light, the last one shows that the power per unit area must decrease as  $r^{-2}$ . In absence of noise the expression of the received waveform is:

$$E_r(f, t, u) = \frac{\alpha(\theta, \psi, f) \cos 2\pi f(t - r/c)}{r}. \quad (7.3)$$

In this case  $\alpha(\theta, \psi, f)$  is the product of the antenna patterns of transmit and receive antennas in the given direction. This takes into account also the changes of placing a receive antenna in the point  $u$ .

Define:

$$H(f) := \frac{\alpha(\theta, \psi, f)e^{-j2\pi fr/c}}{r}. \quad (7.4)$$

$H(f)$  is the transfer function of the LTI system channel. The received field at  $u$  in response to a weighted sum of the transmitted waveforms is simply the weighted sum of responses to those individual waveforms. In the same manner:  $E_r(f, t, u) = \Re[H(f)e^{j2\pi fr/c}]$ .

The second example is a fixed transmit antenna and a receive antenna that is moving with speed  $v$  in the direction of increasing distance from the transmit antenna. Define the same point  $u$  of the first example with  $r(t)=r_0+vt$  in place of  $r$ .

Substituting  $r$  with  $r(t)$  and grouping the terms in the cosine by  $t$  the new electric field at the moving point  $u$  becomes:

$$E(f, t, (r_0 + vt, \theta, \psi)) = \frac{\alpha_s(\theta, \psi, f) \cos 2\pi f((1 - v/c)t - r_0/c)}{r_0 + vt}. \quad (7.5)$$

In the same manner the received waveform becomes:

$$E_r(f, t, (r_0 + vt, \theta, \psi)) = \frac{\alpha(\theta, \psi, f) \cos 2\pi f((1 - v/c)t - r_0/c)}{r_0 + vt}. \quad (7.6)$$

To derive the new frequency is applied a *Doppler shift* of  $-fv/c$  at the old frequency, due to the motion of the observation point. This channel cannot be represented as an LTI channel. The only way to transform it into a LTI one is to ignore the time-varying attenuation in the denominator of the equations (7.5) and (7.6). The transfer function is then (7.4) translated in the frequency by the Doppler shift of  $-fv/c$ .

The analysis above is equivalent if the antennas are inverted: fixed receive antenna and moving transmit antenna.

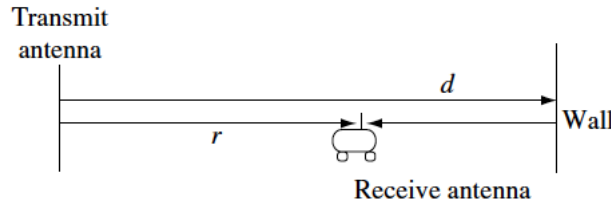


Figure 7.6: Direct path and reflected path from the transmitter to the receiver.

The last example is shown in figure 7.6 and considers two different situations with a large reflecting wall transmitting the same sinusoid  $\cos 2\pi ft$ . The first situation considers both the transmit and the receive antenna fixed. Effectively the waveform received can be calculated by the sum of the free space wave from the transmitter plus the reflected free space waves from each of the reflecting obstacles with a good approximation. A large wall guarantees that the incident and the reflected angle are equivalent, so that the reflected wave from the wall has the intensity of a free space wave at a distance equal to the distance to the wall and then back to the receive antenna:  $2d-r$ . From (7.3) and assuming the same antenna gain  $\alpha$  for both the direct and reflected wave:

$$E_r(f, t) = \frac{\alpha \cos 2\pi f(t - r/c)}{r} - \frac{\alpha \cos 2\pi f(t - (2d - r)/c)}{2d - r}. \quad (7.7)$$

The phase difference between the two waves is:

$$\Delta\theta = \left( \frac{2\pi f(2d - r)}{c} + \pi \right) - \left( \frac{2\pi fr}{c} \right) = \frac{4\pi f}{c}(d - r) + \pi \quad (7.8)$$

The two waves interfere constructively if the phase difference is an integer multiple of  $2\pi$ , destructively if the phase difference is an odd integer multiple of  $\pi$ .

The second situation considers the receive antenna moving at a velocity  $v$ . The strength of the received signal increases until a phase difference of a multiple of  $2\pi$  and then decreases until a phase difference of an odd multiple of  $\pi$  in a periodic cycle. Defining  $r=r_0+vt$  the received waveform is:

$$E_r(f, t) = \frac{\alpha \cos 2\pi f[(1 - v/c)t - r_0/c]}{r_0 + vt} - \frac{\alpha \cos 2\pi f[(1 - v/c)t + (r_0 - 2d)/c]}{2d - r_0 - vt}. \quad (7.9)$$

In this case there are two different Doppler shifts respectively for the first and second term of the equation (7.9). The first sinusoid has a frequency of  $f(1-v/c)$  and the second sinusoid has a frequency of  $f(1+v/c)$ . The first and the second shifts are respectively  $D_1=-fv/c$  and  $D_2=fv/c$ . The difference of these two terms is called Doppler spread:

$$D_s := D_2 - D_1 \quad (7.10)$$

### Wireless Channel Model

The examples above show that on the response to the sinusoidal input  $\phi(t) = \cos 2\pi ft$  the received signal can be written as  $\sum_i a_i(f, t)\phi(t - \tau_i(f, t))$ . The  $a_i(f, t)$  are the product of the attenuation factors, on path  $i$ , due to the antenna pattern of the transmitter and the receiver, the nature of the



reflector and a factor that is function of the distance from the transmitting antenna to the reflector and from the reflector to the receive antenna.  $\tau_i(f, t)$  is the propagation delay at time  $t$  from the transmitter to the receiver on path  $i$ .

Assuming that  $a_i(f, t)$  and  $\tau_i(f, t)$  do not depend on the frequency  $f$  (in practice the attenuations and the propagation delays are usually slowly varying functions of frequency), the above relation becomes:

$$y(t) = \sum_i a_i(t) x(t - \tau_i(t)). \quad (7.11)$$

Where  $y(t)$  is the output of this new model and  $x(t)$  is an arbitrary input with non-zero bandwidth. This model is linear and can be described by the response  $h(\tau, t)$  at time  $t$  to an impulse transmitted at time  $t - \tau$ . Following this way the new input/output relationship becomes:

$$y(t) = \int_{-\infty}^{+\infty} h(\tau, t) x(t - \tau) d\tau. \quad (7.12)$$

Comparing (7.11) and (7.12) the impulse response is:

$$h(\tau, t) = \sum_i a_i(t) \delta(\tau - \tau_i(t)). \quad (7.13)$$

In the case when the transmitter, receiver and the environment are all stationary, the attenuation  $a_i(t)$  and propagation delays  $\tau_i(t)$  do not depend on time  $t$ . The channel is linear time-invariant with an impulse response:

$$h(\tau) = \sum_i a_i \delta(\tau - \tau_i). \quad (7.14)$$

And with a transfer function:

$$H(f, t) := \int_{-\infty}^{+\infty} h(\tau, t) e^{-j2\pi f\tau} d\tau = \sum_i a_i(t) e^{-j2\pi f\tau_i(t)}. \quad (7.15)$$

In this way the relation between transmit and receive antennas is simply represented by an impulse response in the time domain and by a corresponding transfer function in the frequency domain. It's now important to derive a baseband model of the channel, in fact while in typical wireless applications the transmission is given with passband signals, most of the processing is done at the baseband (coding/decoding, modulation/demodulation, synchronization). The transmitter has to transform the baseband signal into a passband one before the transmission; the receiver then down-converts the signal to a baseband one before processing.

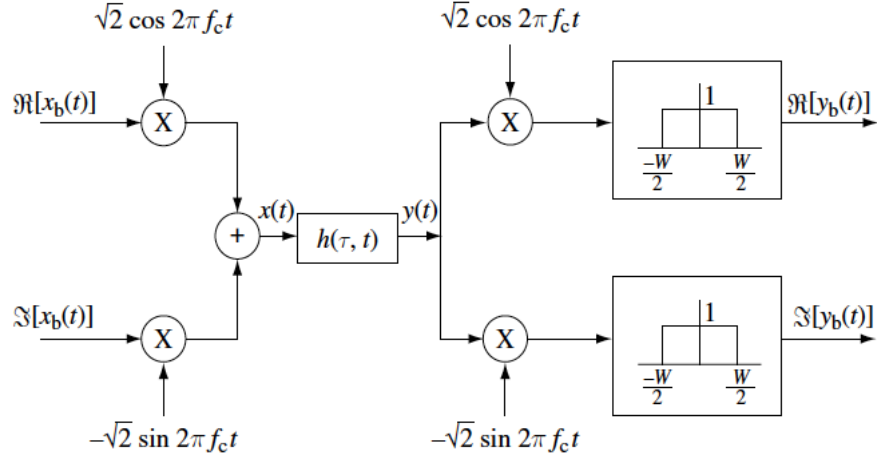


Figure 7.7: System diagram from the baseband transmitted signal  $x_b(t)$  to the baseband received signal  $y_b(t)$ .

Let  $x_b(t)$  and  $y_b(t)$  be the complex baseband equivalents of the transmitted signal  $x(t)$  and the received signal  $y(t)$  (for the calculation of a baseband signal see [8, pp. 22-23]). Figure 7.7 shows the system diagram of a passband communication system known as *quadrature amplitude modulation* (QAM). Substituting  $x(t) = \sqrt{2}\Re[x_b(t)e^{j2\pi f_c t}]$  and  $y(t) = \sqrt{2}\Re[y_b(t)e^{j2\pi f_c t}]$  into (7.11):

$$\begin{aligned}\Re[y_b(t)e^{j2\pi f_c t}] &= \sum_i a_i(t)\Re[x_b(t - \tau_i(t))e^{j2\pi f_c(t - \tau_i(t))}] \\ &= \Re\left[\left\{\sum_i a_i(t)x_b(t - \tau_i(t))e^{-j2\pi f_c \tau_i(t)}\right\}e^{j2\pi f_c t}\right].\end{aligned}\quad (7.16)$$

Similarly:

$$\Im[y_b(t)e^{j2\pi f_c t}] = \Im\left[\left\{\sum_i a_i(t)x_b(t - \tau_i(t))e^{-j2\pi f_c \tau_i(t)}\right\}e^{j2\pi f_c t}\right].\quad (7.17)$$

Hence, the baseband equivalent channel is:

$$y_b(t) = \sum_i a_i^b(t)x_b(t - \tau_i(t)).\quad (7.18)$$

where

$$a_i^b := a_i(t)e^{-j2\pi f_c \tau_i(t)}.\quad (7.19)$$

and the baseband equivalent impulse response is:

$$h_b(\tau, t) = \sum_i a_i^b(t)\delta(\tau - \tau_i(t)).\quad (7.20)$$

The baseband output is the sum, over each path of the delayed replicas of the baseband input.

A discrete-time model can be derived from the continuous-time model by the sampling theorem. The new input/output relationship is:

$$y[m] = \sum_l h_l[m] x[m-l]. \quad (7.21)$$

where:

$$h_l[m] = \sum_i a_i^b(m/W) \text{sinc}[l - \tau_i(m/W)W]. \quad (7.22)$$

and  $W$  is the limited band of the input waveform. When the gains  $a_i^b(t)$  and the delays  $\tau_i(t)$  of the paths are time-invariant, (7.22) can be simplified as:

$$h_l = \sum_i a_i^b \text{sinc}[l - \tau_i W]. \quad (7.23)$$

The channel in this case is LTI and (7.23) can be interpreted as the sample  $(1/W)$ th of the low-pass filtered baseband channel response  $h_b(\tau)$  (see (7.14)) convolved with  $\text{sinc}(W\tau)$ .

Figure 7.8 shows a new diagram of the model: the sampling operation is interpreted as modulation and demodulation in the communication system. At time  $n$  the complex symbol  $x[m]$  is modulated by the sinc pulse before the transmission. The received signal is sampled at times  $m/W$  at the output of the low-pass filter at the receiver.

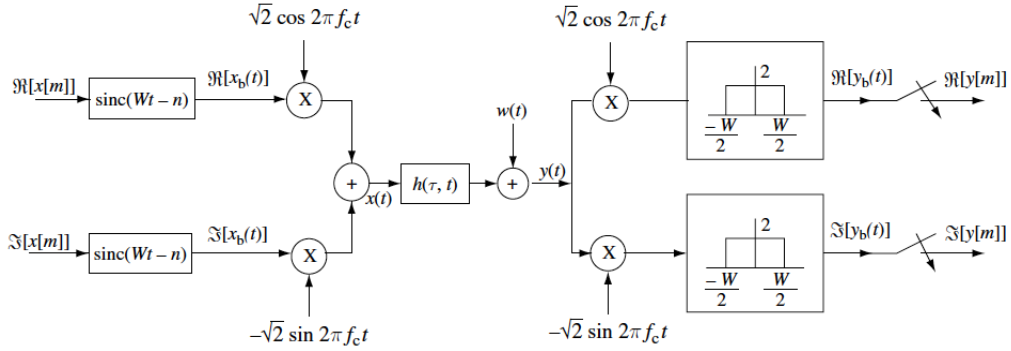


Figure 7.8: System diagram from the baseband transmitted symbol  $x[m]$  to the baseband sampled received signal  $y[m]$ .

As given in the figure, after the convolution of the input  $x(t)$  and the channel impulse response  $h(\tau, t)$  a new term  $w(t)$  is added. It represents the

additive white noise of the system. Considering this terms equations (7.11) and (7.21) becomes:

$$y(t) = \sum_i a_i(t)x(t - \tau_i(t)) + w(t). \quad (7.24)$$

And:

$$y[m] = \sum_i h_l[m]x[m - l] + w[m]. \quad (7.25)$$

Where  $w[m]$  is the low-pass filtered noise at the sampling instant  $m/W$ . The assumption of AWGN is valid if the primary source of the noise is at the receiver or is radiation impinging on the receiver that is independent of the paths over which the signal is being received. This is true for most communication situations.

## Chapter 8

# Conclusions

The first part of this thesis has compared the second generation digital transmission system for cable systems and second generation digital terrestrial television broadcasting system.

A first look at figure 2.1 shows that the T2 system contains many additional blocks as the compensating delay (Input Processing), the constellation rotation (BICM) and MISO processing (OFDM generation) that has been described in the report. On the other side the Data Slice Builder is one of the few blocks that belongs only to the C2-system.

While the input processing is almost the same (Chapter 3), the first differences showed are the code rates of the FEC encoding and the modulations, analyzed in chapter 4.

The frame structure (chapter 6) of the two systems is completely different. In the T2-system the Frame Builder generates superframes, sequences of frames each followed by a future extension frame part. In the C2-system the Frame Builder operates on data slices, groups of PLPs generated by the Data Slice Builder as explained in chapter 5.

Chapter 5 gave also a description of L1 signalling part 2 of the C2-system and a brief analysis of the T2-system signalling, given by P1 and P2 symbols.

The thesis ended with the descriptions of the HFC and the wireless channel model. The first channel is time invariant and, in comparison with the second, has less varying parameters and distortions, as echo, impulse and burst events and Gaussian noise. The wireless channel is typically time-varying and depends on the situation considered. The transmitter and the receiver can be fixed or can moving; as the environment is different in each situation, obstacles like buildings can reflect the signal and generate different paths of the same signal.



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